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OCTOBER 1975

SPECIAL RECEIVER ISSUE

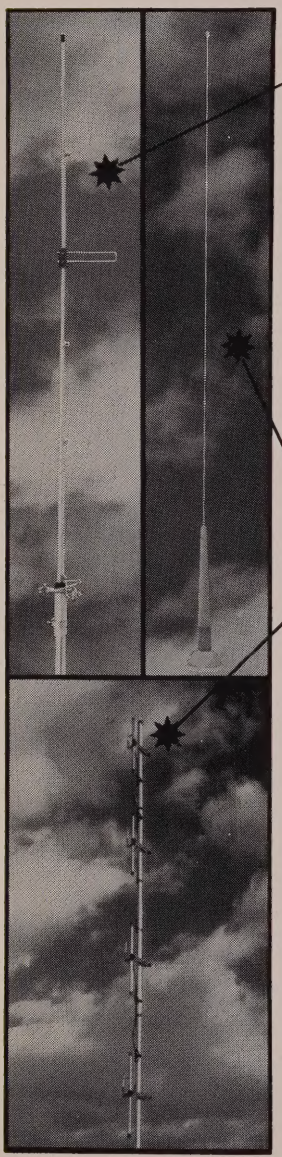


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More Details? CHECK—OFF Page 110

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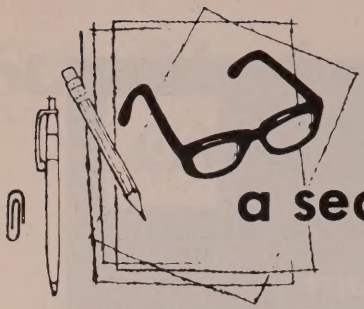
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a second look

by jim fisk

As more and more amateurs are discovering, the commercial double-balanced mixer modules which are currently available are the clear choice when you are designing systems which require frequency conversion. The advantages of using these devices (along with sample circuits) were outlined in *ham radio* as long ago as 1968, and more recently in the other amateur radio magazines. Although the double-balanced mixer offers low distortion and high isolation, it suffers from fairly high conversion loss. While single-balanced mixers offer lower conversion loss and can be built less expensively, they do not provide the low distortion and high isolation characteristics of the double-balanced design.

Now, however, engineers at Hewlett-Packard have developed a single-balanced mixer with isolation, distortion and conversion-loss characteristics that equal or exceed those of some double-balanced mixers now on the market. Furthermore, the H-P printed-circuit balanced mixers, part number 5082-9200, are many times smaller (0.57" wide, 0.50" high).

With local-oscillator injection in the range from 200 to 900 MHz, conversion loss of the H-P balanced mixer is about 6.5 to 7.5 dB, a nearly 1 dB improvement over the performance of double-balanced circuits. In addition, isolation is 15 to 20 dB higher, being 45 dB at 200 MHz and 25 dB at 900 MHz. The maximum frequency is specified at 1200 MHz although selected devices would very likely provide good performance on the 1296-MHz amateur band.

In terms of distortion, at a local-oscillator level of +10 dBm (10 mW), the third-order intercept is typically +23 dBm and the 1 dB compression point occurs at +6 dBm. The second-order intercept is at +38 dBm, a roughly 10 dB improvement over conventional double-balanced mixers. Although a filter must be added at some later stage to separate the rf and i-f signals, this is not a particular disadvantage in communications circuits because a selective i-f filter is usually placed in the circuit immediately after the frequency-conversion stage.

Although balanced mixers are never quite as good in practice as they are in theory because of circuit imbalances, the various mismatches often cancel each other to a large degree. To achieve high performance in the new H-P mixer engineers further improved the matching, balancing and symmetry of the circuitry by removing the human element from the manufacturing process. The transformers, for example, are not wound by hand, but are etched on a printed-circuit board with spacing between windings controlled down to 0.001 inch (0.03mm). This technique also balances the parasitic capacitances between windings. With hand winding these elements vary widely as does the overall balance of the circuit.

Since diode matching is important in balancing a mixer, the H-P designers chose their own Schottky process because of its high yield of matched-diode pairs. With the carefully controlled printed-circuit transformers and closely matched diodes, the result is a 1% matching in forward voltages and a 0.1 pF capacitance match.

Although priced at $11 in small quantities, the cost of the H-P mixer drops to less than $2 each in production quantities, so it should see widespread use in commercial applications.

Jim Fisk, W1DTY
editor-in-chief

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More Details? CHECK–OFF Page 110

presstop

THE GETTYSBURG LOG-JAM continues to plague both Amateurs and the FCC as Amateur license applications now are taking up to 12 weeks for final action. This is true even after special efforts (HR Report, August 1) to dig out. Unfortunately, no Amateur tickets were processed while the attempt to catch up on CB applications was underway.

Expansion Plans Provide Hope for the future. Included are a larger, faster computer and a 25% increase in personnel. Helping already are improved procedures and the new box numbers. Remember, address your Amateur applications to Box 1020, Gettysburg, Pennsylvania 17325.

SIX-METER AMATEUR BAND may be in trouble according to Prose Walker's comments at the Texas VHF FM Society's convention. Prose seems to feel that there's a definite push on to add channel 1 to the VHF TV spectrum, and channel 1 is 6 meters!

The Channel 1 agitation may be an offshoot of a proposal made earlier this year by a Dallas consulting firm in response to Docket 20264, which pertains to radio callbox systems in the 72-74 MHz range, suggesting that those four MHz be added to two MHz from the six-meter band to create a new TV channel.

RTTY ENTHUSIASTS are warned that parts for Teletype Models 14, 15, 19 and 20 are going to become harder to get in the near future. Teletype Corp. has announced they'll stop accepting orders for all replacement parts for those models on December 1, 1975, so better order your spares now.

Questions Or Orders can go to Teletype, 5555 Touhy Avenue, Skokie, Illinois 60076, Attn: R.A. Morton, Dept. 3121 — phone (312) 982-2168. Thanks to W8KAJ.

AMSAT QSL BUREAU MANAGER WA1EHF has made a sudden move to California and plans to move the QSL bureau out there as soon as he has a permanent address. In the meantime, WA1EDX will be filling in the void for Dennis and the 288 Grand Street, Bridgeport, Connecticut 06604 address is still valid.

Satellite Use for location of downed aircraft has been proposed, and AMSAT has been asked to participate in feasibility studies. Anyone interested — particularly members in the mid-Atlantic (Washington) area — should contact AMSAT.

OSCAR AND THE HAM, a new video tape copyrighted by the ARRL, has been released for viewing on local PBS stations. Anyone interested should contact their local station and request that they run it.

Help Is Needed in making video tape copies. If you can give any assistance, please get in touch with ARRL Headquarters.

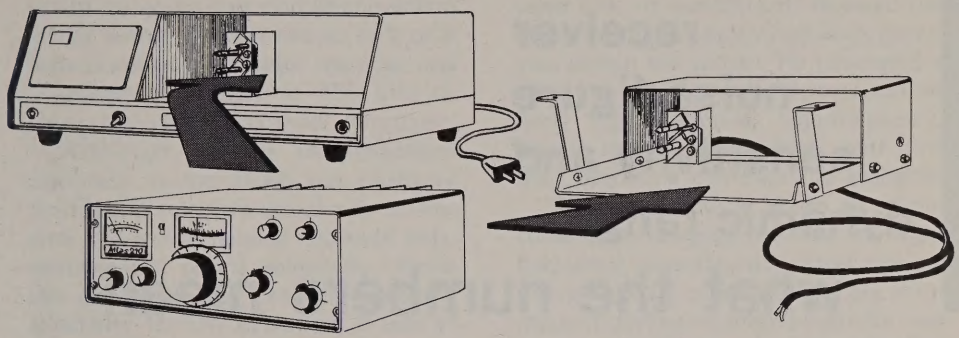
"HOT" HAM GEAR does show up and it pays to be able to prove positive ownership when it does. WB9EBO found his Standard 826 and synthesizer stolen from his home last spring for sale at Hamfesters' hamfest near Chicago, kept an eye on them while W9JZK called the Cook County Sheriff's Police for assistance. Gerry's call, electric pencilled inside the chassis, left the seller (a non-ham who had "bought them from some guy at another hamfest") without an argument.

EUROPEAN BOUND U.S. HAMS should seriously consider taking along a hand-held tuned up for appropriate repeaters on the low end of two meters. W9JUV heard lots of activity in Germany and England on a tunable receiver during a two-week trip in August, and temporary operating permits for many parts of Europe are much easier to get than are their equivalents here.

R. L. DRAKE COMPANY FOUNDER and president Bob Drake passed away July 28th after a long illness. Bob, W8CYE, founded the R.L. Drake Company in 1942 and was active in the successful business until recently. His son Peter has been elected president and promises to continue the traditions of this Amateur-oriented firm. No policy or personnel changes are anticipated in the near future.

JOSEPH JOHNSON, W3GGO, has been named Chief, Rules and Legal Branch of the Amateur and Citizens Division of the FCC effective August 17. Joe replaces John Johnston, K3BNS. This is an important position and we are glad to again see an Amateur in this chair.

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receiver noise figure sensitivity and dynamic range — what the numbers mean

A complete discussion of receiver sensitivity, intermodulation distortion, cross modulation and gain compression, and what they mean in terms of performance

James R. Fisk, W1DTY, Ham Radio Magazine, Greenville, New Hampshire 03048

When it came to receivers, the earliest amateur operators were concerned primarily with sensitivity and experimented almost endlessly with different types of crystals, trying to find the one that was the most sensitive. Then came DeForest's Audion, and Armstrong's regenerative detector, and amateurs who could afford the tubes found they had all the sensitivity they could use. However, as the hobby grew, and more and more amateurs started populating the band below 200 kHz, the simple regenerative detector simply wasn't up to the task. Selectivity, with simple tuned input circuits, was practically nonexistent, and the regenerative detector hopelessly overloaded in the presence of strong signals.

In the early 1920s amateurs worked to improve their tuners, but even the so-called "low-loss" tuners were only marginally acceptable. Although several superheterodyne designs were described in the amateur magazines, it wasn't until low-cost, commercial i-f transformers became available in the late twenties that the superhet saw widespread amateur use. Selectivity against interfering signals was still a problem, however, and James Lamb revolutionized receiver design in 1932[1] with his "single-signal" CW circuit which used an i-f stage with extremely high selectivity — provided by regeneration or a simple crystal filter.

The single-signal, single-conversion superhet of the late 1930s suffered from poor rf image response at the higher frequencies, but it wasn't too severe on 14 MHz and few amateur receivers of the day, in fact, tuned much above 18 or 20 MHz (15 meters was not yet assigned to amateur use and most 10-meter operators used specialized receivers or converters). When the 10- and 15-meter bands opened up after the war, however, the poor rf image response of the

single-conversion superhet with a 455-kHz i-f had to be faced — it was solved by going to a double-conversion layout with a first conversion to 2 or 3 MHz to minimize rf image response, and a second conversion to 455 kHz or lower for adjacent channel selectivity.

Although amateur radiotelephone operation in the 1930s was relatively limited, the huge growth of a-m activity after the war demanded improved adjacent-channel phone selectivity. While the crystal filter provided excellent selectivity for CW operation, it was of little or no use on a-m or ssb and some phone operators started using a Q5er — an outboard 80-kHz i-f strip — for improved phone selectivity. This led to the triple-conversion superhets which were the rage of the 1950s.

As pointed out by Goodman,[2] however, the multiple-conversion design had many shortcomings, including the large number of stages between the antenna and the high-selectivity i-f which made it practically impossible to attenuate strong, adjacent signals. And, with at least three oscillators running at the same time, it was difficult to avoid the many spurious signals which were generated within the system. He advocated a return to the single-conversion superhet using highly-selective, high-frequency, crystal lattice filters which were just then becoming commercially available.

The 1960s saw a return to single-conversion designs, the use of high-frequency, crystal-lattice filters and the widespread use of semiconductors. With modern devices receiver sensitivity was no longer limited by the rf amplifier (or mixer) stage, but by the external galactic and man-made noise. Cross modulation and overload, on the other hand, were becoming a serious problem as more and more amateurs started using high-power linears and large, directive antennas. Modern communications receivers, therefore, in addition to meeting stringent frequency accuracy, stability, sensitivity, and selectivity requirements, must provide freedom from cross modulation, intermodulation distortion and blocking. Some modern, solid-state solutions to these design goals were discussed recently by Rohde.[3]

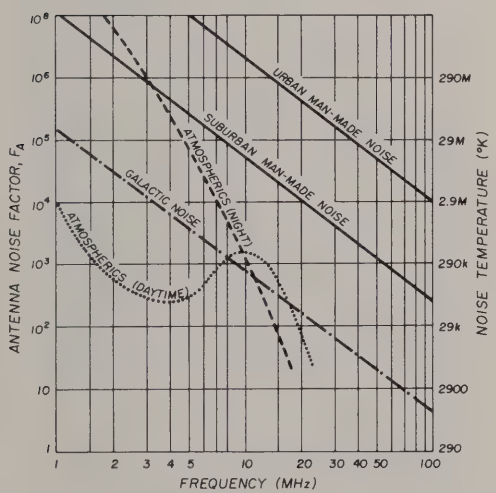


fig. 1. Receiver sensitivity is limited by the external available noise power which varies with frequency. For quiet, rural locations galactic noise is the limiting factor down to about 18 MHz, and atmospheric noise dominates below 18 MHz.

The specifications for a typical modern communications receiver might list sensitivity of 0.5 μV for 10 dB signal-plus-noise-to-noise (S+N/N) ratio, intermodulation distortion of -65 dB, "wide" dynamic range, and "virtual elimination" of overload from adjacent signals.

However, is 0.5 μV sensitivity for 10 dB S+N/N adequate for operation on the high-frequency bands? For satellite communications on 10 meters? What is -65 dB intermodulation distortion in terms of signal strength? "Wide dynamic range" and "virtual elimination of overload" are obviously advertising superlatives without definition but what, exactly, can you expect from a high-quality, modern receiver design? Perhaps, if these performance data were defined,

and amateurs understood what they meant, manufacturers would be encouraged to use no-nonsense numerical data. Only then can amateurs compare the dynamic range and cross-modulation performance of one receiver against that of another.

sensitivity

The minimum usable signal or sensitivity of a receiver is determined by the noise in the receiver output. This can be noise generated within the receiver, thermal noise generated by losses in the transmission line, or atmospheric, man-made or galactic noise picked up by the antenna. As shown in **fig. 1**, external noise sources are likely to be the limiting factor up to 100 MHz or so.[4] In urban areas man-made noise predominates and measurements indicate the average level of man-made noise in suburban areas is about 16 dB lower. In a quiet, rural location which has been chosen with care the man-made noise may be near the galactic noise level, but few amateurs are so fortunate.

Atmospheric noise usually predominates in quiet locations at frequencies below about 20 MHz and is produced by lightning discharges so the level depends upon a number of variables including frequency, weather, time of day, season and geographical location. This type of noise is particularly severe during rainy seasons near the equator and generally decreases at the higher latitudes. More complete data on high-frequency atmospheric noise is given in reference 5.

Galactic or cosmic noise is defined as rf noise caused by disturbances which originate outside the earth or its atmosphere. The primary causes of this noise, which extends from 15 MHz well into the microwave region, are the sun and a large number of noise sources distributed chiefly along the Milky Way. Solar noise can vary as much as 40 dB from "quiet" sun levels (low sunspot activity) to periods of "disturbed" sun (high sunspot activity). Galactic noise from the center of the Milky Way is about 10 dB below the noise from a "disturbed" sun, whereas noise levels from other parts of the galaxy can be as much as 20 dB lower. This is important in satellite communications and will be discussed later.

thermal noise

The free electrons in any conductor are in continuous motion — motion that is completely random and is the result of thermal agitation. The effect of this electron motion is to cause minute voltages which vary in a random manner to be developed across the terminals of the conductor. Since this phenomenon was first demonstrated by J. B. Johnson in 1928,[6] thermal noise is sometimes known as Johnson noise. At the same time, H. Nyquist showed, on the basis of the statistical theory of thermodynamics, that the mean square noise voltage generated in any resistance can be expressed as [7]

$$e^2 = 4kTBR \qquad \textbf{(1)}$$

where e^2 = mean square noise voltage
k = Boltzmann's constant = 1.38×10^{-23} joules/°K
T = absolute temperature, °K
B = bandwidth, Hz
R = resistance, ohms

Note that the noise voltage is dependent upon the bandwidth across which it is measured. This implies that noise is evenly distributed across all frequencies which, for all practical purposes, it is.* Although noise bandwidth is not precisely the same as the 3-dB bandwidth of a receiver, in modern receivers with high skirt selectivity the 3-dB bandwidth can be used in **eq. 1** with little error.

The equivalent circuit of any impedance as a source of noise voltage is

*At extremely high frequencies statistical mechanics is no longer valid, and **eq. 1** must be revised on the basis of quantum theory. This equation is valid, however, to at least 6000 GHz.[8]

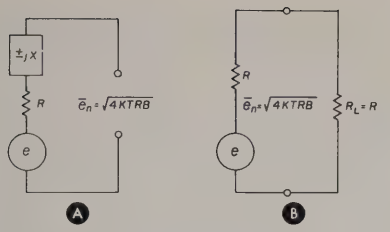


fig. 2. Mean noise voltage depends on temperature, resistance and bandwidth, and is completely independent of reactance as shown in (A). Maximum noise power is transferred to the load when the load resistance is matched to the source resistance (B).

shown in **fig. 2A**. Note that the thermal noise voltage is dependent only on the resistive component and is independent of any reactance in the circuit. As might be expected, maximum noise power is transferred from a thermal noise source when the load impedance presents a conjugate match to the source impedance. This is represented in **fig. 2B** where the load resistance, R_L, is equal to the source resistance. Since $R = R_L$, the noise voltage developed across the load is $e/2$, and from Ohm's law

$$P = \frac{E^2}{R} = \frac{(e/2)^2}{R} = \frac{e^2}{4R} \text{ watts} \quad \textbf{(2)}$$

Substituting the value of e^2 from **eq. 1** into **eq. 2**, the power which can theoretically be transferred under such conditions is called the *available noise power* and is given by

$$P_n = kTB \quad \textbf{(3)}$$

The factor of $4R$ has cancelled out so the available noise power does not depend upon the value of the resistance. This is significant because it means that the available noise power of any resistor (or any noise source), if measured over the same bandwidth, can be represented by a resistor at temperature T. Thus, every noise source has an equivalent noise temperature.

The actual noise power dissipated in the load resistance may be affected by loss in the connecting leads, noise power generated in the load resistor itself, or a less than perfect match to the original resistance. This property is sometimes used in low-noise uhf amplifiers by creating a deliberate (but carefully determined) mismatch between the input termination and the detection device so that something less than the available termination noise power is coupled into the detector.

signal-to-noise ratio and receiver noise figure

The relation of signal amplitude to noise is commonly referred to as the signal-to-noise (S/N) ratio. Unfortunately, this ratio has not been well standardized and is often used interchangeably to mean the ratio of rms signal voltage to rms noise voltage, the ratio of peak signal voltage to peak noise voltage and, in pulse systems, the ratio of peak signal power to average noise power. Therefore, when discussing S/N ratio, it's important to determine exactly which ratio is being referred to.

Although the minimum discernible signal (MDS) that can be heard above the receiver noise level is sometimes used as an indication of receiver sensitivity, it is extremely subjective because it differs many dB from measurement to measurement, and from one operator to another (some experienced weak-signal operators can detect signals as much as 20 dB below the noise level while other operators may have difficulty discerning signals which are equal to the noise level).[9]

Receiver sensitivity has also been defined in terms of a signal-to-noise ratio of unity (signal equals noise)* or equivalent noise floor, but this is difficult to measure unless you have a calibrated signal generator and a spectrum analyzer. Noise figure or noise factor, on the

*This is sometimes erroneously referred to as *tangential* sensitivity. Tangential sensitivity, however, corresponds to a signal-to-noise ratio of 6.25 and is about 8 dB higher.[10]

other hand, is less susceptible to measurement errors than sensitivity and, since its introduction in 1944 by Friis,[11] it has become the accepted figure of merit for receiver sensitivity. Noise figure, NF, is simply noise factor, F, expressed in dB.

$$NF = 10 \log F\,(dB) \qquad (4)$$

The concept of noise factor allows the sensitivity of any amplifier to be compared to an ideal (lossless and noiseless) amplifier which has the same bandwidth and input termination. As far as noise is concerned, that part of a receiver between the antenna and the output of the i-f amplifier can be regarded as an amplifier. The fact that the mixer stage shifts the frequency of the noise does not change the situation — it merely causes the noise to lie in a different place in the spectrum from the input noise. The only exception is when the receiver has poor rf image rejection. In this case the noise figure of the receiver is 3 dB worse than it would be if the same receiver had good rf image rejection because the image noise appears at the output along with noise associated with the desired received frequency. This effectively doubles the noise at the output of the i-f amplifier.*

The noise factor, F, of a receiver is defined as

$$F = \frac{S/N\,(ideal\ receiver)}{S/N\,(practical\ receiver)} = \frac{S_i/N_i}{S_o/N_o} \qquad (5)$$

S_i = available signal input power
N_i = available noise input power
S_o = available signal output power
N_o = available noise output power

Using this definition, it can be seen that an ideal receiver adds no noise to a signal so its output signal-to-noise ratio is the same as that at the input and the noise factor, $F = 1$.

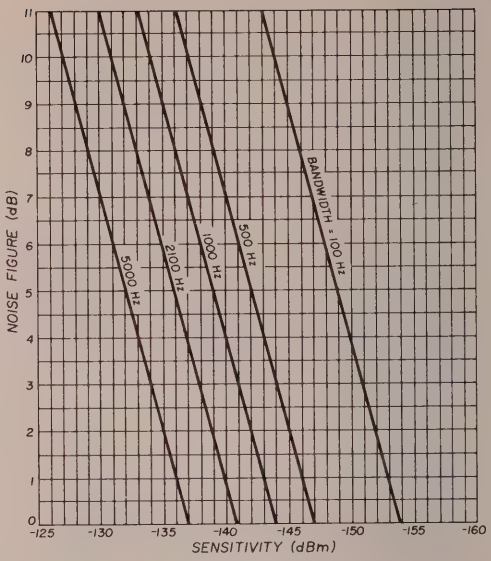


fig. 3. Receiver sensitivity (−dBm) vs receiver noise figure and bandwidth for unity signal-to-noise ratio (signal equals noise). Add 10 dBm for a S/N ratio of 10 dB. Table 1 lists microvolt sensitivity for 50- and 75-ohm systems in terms of dBm sensitivity.

Since the available noise input power, N_i is defined as kT_oB in **eq. 3**, and the power gain of the system, $G = S_o/S_i$, **eq. 5** can be rewritten as

$$F = \frac{N_o}{GkT_oB} \qquad (6)$$

Where T_o is 290° kelvin (IEEE definition). With the receiver noise factor defined in terms of noise output power, N_o, power gain, G, and noise input power, kT_oB, noise factor can be easily correlated to receiver sensitivity. Consider the case where the output signal-to-noise ratio, S_o/N_o, is unity

$$S_o = N_o$$

*The noise figure is always defined at the *input* of the final detector (i-f output) because the noise output of a detector (but not of a mixer) is affected by the presence of a signal. An fm signal, for example, will suppress weak noise but will be suppressed itself by strong noise.

For this specialized case, **eq. 6** can be rewritten

$$F = \frac{S_i}{kT_oB} \qquad (7)$$

When the temperature is 290°K and the bandwidth is in kHz, $kT_oB = 4 \times 10^{-15}$ mW per kHz. Rewriting **eq. 7** in terms of dBm (dB referenced to 1 mW)

$$S_i = 10 \log kT_o + 10 \log B + NF \qquad (8)$$
$$= -144 + 10 \log B_{kHz} + NF(dBm)$$

This function is plotted graphically in **fig. 3** for bandwidths commonly found in amateur communications receivers. For example, assume a high-frequency receiver has an 8 dB noise figure at 14.2 MHz and a bandwidth of 2.1 kHz. From **eq. 8** or **fig. 3**, the noise floor of the receiver at 14.2 MHz is at about −133 dBm. An input signal of −123 dBm (10 dB greater) would be required for a 10 dB S/N ratio.* To convert dBm to microvolts, recall that

$$E = \sqrt{RP} \qquad (9)$$

where E is in volts, R is resistance in ohms and P is power in watts. Since −123 dBm is 5.01×10^{-16} watts, −123 dBm is equivalent to 0.16 μV *across* a 50-ohm input termination. However, for a matched signal source, as shown in **fig. 4**, where the source resistance, R_s, is equal to the load resistance, R_L, the source voltage must be twice the voltage across R_L because of the voltage-dividing effect of the two series resistors in the network. For a *matched* 50-ohm source, therefore, an input signal of −123 dBm requires a source voltage of 0.32 μV. A chart of dBm vs microvolts for *matched* 50- and 75-ohm systems is presented in **table 1.**

*This is the signal-to-noise ratio in ssb and CW reception. The S/N ratio of a-m and nbfm signals is somewhat less because a-m (and nbfm) detection use only the envelope as a useful output and the S/N ratio must be reduced by a factor which is related to percentage of modulation (or modulation index).

table 1. Microvolt sensitivity vs dBm for *matched* 50- and 75-ohm receiving systems.

dBm	μV 50 ohms	μV 75 ohms	dBm	μV 50 ohms	μV 75 ohms
- 76	70.8	86.7	-111	1.26	1.54
- 77	63.2	77.4	-112	1.12	1.38
- 78	56.2	69.1	-113	1.00	1.23
- 79	50.2	61.5	-114	0.90	1.09
- 80	44.8	54.9	-115	0.80	0.97
- 81	39.8	48.7	-116	0.71	0.87
- 82	35.6	43.6	-117	0.63	0.77
- 83	31.6	38.7	-118	0.56	0.69
- 84	28.2	34.5	-119	0.50	0.62
- 85	25.2	30.9	-120	0.45	0.55
- 86	22.4	27.4	-121	0.40	0.49
- 87	20.0	24.5	-122	0.36	0.44
- 88	17.8	21.8	-123	0.32	0.39
- 89	15.8	19.4	-124	0.28	0.35
- 90	14.2	17.3	-125	0.25	0.31
- 91	12.6	15.4	-126	0.22	0.27
- 92	11.2	13.8	-127	0.20	0.25
- 93	10.0	12.3	-128	0.18	0.22
- 94	9.0	10.9	-129	0.16	0.19
- 95	8.0	9.7	-130	0.14	0.17
- 96	7.1	8.7	-131	0.13	0.15
- 97	6.3	7.7	-132	0.11	0.14
- 98	5.6	6.9	-133	0.10	0.12
- 99	5.0	6.2	-134	0.09	0.11
-100	4.5	5.5	-135	0.08	0.10
-101	4.0	4.9	-136	0.071	0.087
-102	3.6	4.4	-137	0.063	0.077
-103	3.2	3.9	-138	0.056	0.069
-104	2.8	3.5	-139	0.050	0.062
-105	2.5	3.1	-140	0.045	0.055
-106	2.2	2.7	-141	0.040	0.049
-107	2.0	2.5	-142	0.036	0.044
-108	1.8	2.2	-143	0.032	0.039
-109	1.6	1.9	-144	0.028	0.035
-110	1.4	1.7	-145	0.025	0.031

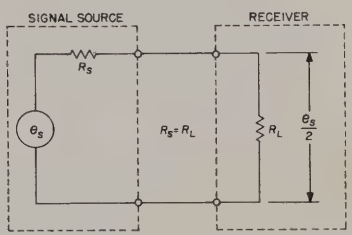


fig. 4. When a signal source is matched to a load, the voltage across the load is one-half the source voltage because of the voltage-dividing effect of the source and load resistors. When making receiver sensitivity measurements in the laboratory, a 6-dB attenuator is placed in the line so sensitivity can be read directly from the signal-generator's internal calibrated attenuator.

Because of this two-to-one voltage-dividing effect, you must be very careful when comparing the sensitivity of one receiver against that of another. An *input* of +119 dBm, for example, implies an input directly at the receiver terminals and is 0.25 μV rms across 50 ohms. Sensitivity, on the other hand, implies the use of a matched signal generator so *sensitivity* of 0.25 μV corresponds to an *input* of −125 dBm. This is a 6 dB difference. Since most amateur receiver manufacturers tend to use *sensitivity* specifications, there is no advantage, but the difference must be considered when you calculate the receiver noise figure and dynamic range. In this article inputs will be stated in dBm as this eliminates the 6 dB conversion factor — **table 1** can be used to convert to *sensitivity*.*

cascaded stages

A relatively simple equation for the noise factor of a receiving system, in terms of the individual stage gains and noise factors is

$$F_T = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} \qquad \textbf{(10)}$$

F_T = overall system noise factor
F_1 = noise factor of the first stage
F_2 = noise factor of the second stage
F_3 = noise factor of the third stage
G_1 = power gain of the first stage
G_2 = power gain of the second stage
G_3 = power gain of the third stage

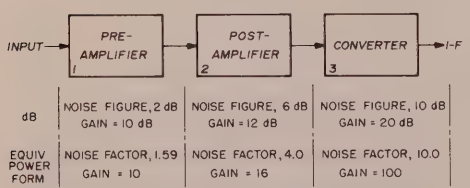


fig. 5. Noise factor of cascaded stages can be calculated by using equation 10 where all quantities are in power ratios. Noise factor at the input of the three cascaded stages shown here is 1.946 (noise figure = 2.9 dB).

Note that all terms are in power ratios. If the gain of the first stage is high, and the noise factor of the second stage is low, the overall system noise factor is determined primarily by the first stage and the third term of **eq. 10** may be dropped.

For example, consider the block diagram of the first three stages of a typical vhf receiving system shown in **fig. 5.** The overall system noise factor is

$$F_T = 1.59 + \frac{4 - 1}{10} + \frac{10 - 1}{10 \cdot 16}$$

$$= 1.59 + 0.3 + 0.056 = 1.946$$

$$NF_T = 10 \log 1.946 = 2.9 \text{ dB}$$

For this receiver the overall system noise figure is 0.9 dB higher than that of the preamplifier alone. Raising the preamplifier gain to 13 dB or dropping the noise figure of the second stage to 4 dB ($F = 2.5$) would reduce the system noise figure to approximately 2.45 dB ($F = 1.75$). Depending on the frequency and the application, this may be a worthwhile improvement.

noise temperature

It is often convenient when working with very low noise uhf and microwave receivers to represent the noise figure of the receiver as an equivalent noise temperature. This is because the noise temperature of a receiving system varies over the range from 0° to 174°K as the noise factor varies from 1.0 to 1.6 (0 to 2 dB noise figure) and noise calculations using equivalent noise temperatures will provide better accuracy.

As mentioned above, the noise figure of an ideal receiver is 1, so the component of receiver noise figure which is due to internally generated noise, N_r, is

*The usual procedure for measuring receiver sensitivity is to place a 6 dB attenuator between the signal generator and the receiver. Receiver sensitivity can then be read directly from the signal generator's calibrated output attenuator.

$F - 1$. Therefore

$$N_r = (F - 1) GkT_o B \qquad (11)$$

The internally generated noise, N_r, can be represented by a noise power, GkT_rB, where T_r is the equivalent noise temperature. Substituting into **eq. 11**, this equivalent noise temperature can be expressed in terms of the reference noise temperature, T_o = 290°K.

$$T_r = (F - 1) T_o \qquad (12)$$

Eq. 12 can be easily rearranged to express noise factor as a function of the receiver noise temperature, T_r, and the reference noise temperature, T_o

$$F = 1 + \frac{T_r}{T_o} \qquad (13)$$

Although noise temperature is seldom used by amateurs, it is a more basic unit than noise factor and is actually easier to deal with, both in understanding concepts and making practical noise calculations. For a more complete discussion of noise temperature, see reference 12.

transmission lines

The transmission line's contributions to receiver noise come from a common source: line losses. The first of these is the more obvious. When a signal travels down a lossy transmission line, the signal is attenuated. This reduces the signal-to-noise ratio and is equivalent to increasing the noise factor of the receiver. This increase can be calculated by introducing a loss factor, L, which is the loss of the cable expressed as a power ratio.

The second effect is due to the noise factor of the transmission line. The fact that the line has losses implies that there is a loss resistance associated with it (which is distinct from characteristic impedance). Since the line is warm it generates noise due to thermal agitation. The noise factor of the line, F_t, is related to the loss factor, L, and the physical temperature of the line, T_t, by the following equation

$$F_t = \frac{(\frac{1}{L} - 1) T_t}{290} + 1 \qquad (14)$$

The degradation of the receiver noise figure due to transmission line contributions may be calculated by considering the transmission line and the receiver as cascaded stages and using a form of **eq. 10**

$$F_{tr} = F_t + \frac{F_r - 1}{L} \qquad (15)$$

where F_{tr} = noise factor of the receiver and transmission line
F_r = noise factor of the receiver
F_t = noise factor of the transmission line
L = loss factor of the transmission line

For example, a receiver with a noise factor of 4 (NF = 6 dB) is used with a transmission line which has a loss factor of 0.63 (2 dB loss). The physical temperature of the line is 300°K (F_t = 1.61). What is the combined noise figure?

$$F_{tr} = 1.61 + \frac{4 - 1}{0.63} = 6.37$$

$$NF_{tr} = 8.04\ dB$$

When line losses are low but receiver noise figures are 3 dB or greater, line loss is the predominate contributor to increased noise figure. When receiver noise figure is very low, the thermal effect predominates.

antenna noise

Of all the contributions to system noise, antenna noise is probably the least understood. Assuming the antenna is built of good conducting materials, it contributes virtually no thermal noise of its own to the receiving system. The noise power the antenna does deliver to the receiver depends almost entirely on

table 2. Performance of a receiver with 0.5 μV sensitivity for 10 dB S+N/N with 100 feet (30.5m) of RG-8A/U transmission line is shown in first two columns. Third column lists external available noise power for quiet receiving locations on each of the amateur bands. Fourth column shows receiver signal (50-ohms) required for 10 dB S+N/N on each of the bands (based on external noise). Last column lists acceptable noise figure for each of the bands (see text). Bandwidth = 2.1 kHz.

frequency	noise factor at antenna	noise figure	external available noise power	receiver input signal for 10 dB S+N/N	acceptable noise figure
1.8 MHz	15.8	12.0	- 93 dBm	15.3 μV	45 dB
3.5 MHz	16.2	12.1	-101 dBm	12.6 μV	37 dB
7.0 MHz	16.7	12.2	-111 dBm	4.0 μV	27 dB
14.0 MHz	17.6	12.5	-113 dBm	3.1 μV	24 dB
21.0 MHz	18.3	12.6	-118 dBm	1.8 μV	20 dB
28.0 MHz	18.9	12.8	-123 dBm	1.0 μV	15 dB
50.0 MHz	20.9	13.2	-129 dBm	0.5 μV	9 dB
144.0 MHz	26.9	14.2	-139 dBm	0.2 μV	2 dB

the temperature and other physical characteristics of the material lying in the antenna's field of view. A 432-MHz moonbounce receiving antenna looking out into space, for example, may deliver only as much noise power as a resistor at 10°K (noise factor = 1.03). If this same antenna is rotated so that the warm earth comes into its field of view, the antenna noise temperature would rise to about 300°K (noise factor = 2).

A ten-meter Oscar receiving antenna which is pointed into "cold" space, on the other hand, will see a noise temperature of about 10,000°K minimum (noise factor = 35). Pointed at the horizon, however, the antenna noise temperature may be ten or fifteen times higher, depending upon the amount of man-made noise.

There is little that can be done to improve the situation of a terrestial radio circuit, but an antenna that looks at the sky, such as a satellite antenna, deserves careful design. This is because the effect of the earth is still present, and any sidelobe that sees the earth will pick up thermal noise. This is sometimes quite serious and sidelobes are of major concern in many deep-space communications and radio astronomy systems. Careful attention to antenna design with respect to sidelobes can provide antenna temperatures significantly under 50°K, while poor design can result in much higher values.

minimum usable sensitivity

With an understanding of receiver noise factor and its relationship to signal-to-noise ratio, it's now possible to determine the *minimum usable sensitivity* (MUS) of a receiving system, and how the performance of your own equipment affects your ability to receive weak signals. Let's first consider that modern communications receiver mentioned earlier which had a specified sensitivity of 0.5 μV for 10 dB S+N/N ratio.*

From **table 1** a 0.5 μV *sensitivity* for 10 dB S+N/N is equivalent to a sensitivity of -119 dBm in a 50-ohm system (-129 dBm noise floor). Assuming a bandwidth of 2.1 kHz, the noise figure of the receiver is about 11.8 dB (noise factor = 15.1). Assuming 100 feet

*A S+N/N ratio of 10 dB corresponds to a S/N ratio of 9.54 dB, a difference small enough to be neglected for all practical purposes. A S+N/N ratio of 3.5 dB, however, corresponds to a S/N ratio of about 0.9 dB, and the difference must be considered. S+N/N may be converted to S/N by using the relationship [(S+N/N) – 1] = S/N (in power ratios, *not* dB).

(30.5m) of RG-8A/U transmission line at 300°K, the noise factor at the antenna terminals may be calculated from **eq. 15**, and is shown in **table 2** for the six high-frequency amateur bands, (calculated on the basis of 0.5 μV sensitivity for 10 dB S+N/N on all bands, which may be optimistic). Even at 28 MHz, where the line loss has increased the noise factor by 25 per cent, the system noise factor is still well below the available noise power seen by the antenna (see **fig. 1**).

At 50 MHz, however, the system is limited by receiver noise and a lower noise figure would be required for weak-signal work (system noise at 50 MHz in this example is about 2.4 dB higher greater than external noise for a quiet location). If the receiver was connected directly to the antenna terminals to eliminate transmission line losses the system noise figure would be essentially that of the receiver alone and a 0.5 μV signal would provide the desired 10 dB S+N/N ratio. Although 100 feet (30.5m) of RG-8A/U coaxial cable has only about 1.35 dB loss at 50 MHz, it degrades the noise figure sufficiently that the system is no longer limited by external noise sources. This points up the importance of using low-loss transmission lines (or mounting a receiving preamp at the antenna).

Assuming a quiet, rural location that is limited primarily by galactic noise down to about 18 MHz, and atmospheric noise below 18 MHz, what is the *minimum* usable receiver sensitivity for terrestial communications?

As can be seen from **table 2**, rather poor receiver sensitivity is acceptable on 40, 80 and 160 meters because the external noise at these frequencies is very high. This also explains why the simple receivers of the 1920s were relatively successful. The high external noise levels also make it possible to use rather inefficient receiving antennas on the lower frequencies.[13] It's important to note that a 0.5 μV signal is of little practical use on 160, 80 or 40 meters because it would be buried in the noise level.

The required sensitivity on 20, 15 and 10 is not difficult to obtain with modern devices, but receivers which are optimized for the lower frequencies may not offer top performance on 10 meters. It should be pointed out that the "acceptable" noise figure in the last column of **table 2**, is somewhat arbitrary and is based on setting the receiver noise floor about 3 dB below the external noise floor. This is probably adequate 90 per cent of the time, but since noise varies randomly, a statistical analysis indicates there may be times when a lower noise figure may be desirable. However, it is generally agreed that a 10 dB noise figure is more than adequate up to 22 MHz and an 8-dB noise figure may occasionally prove useful on 10 meters. Why design a high-frequency receiver for extraordinary sensitivity when its performance is limited by external noise over which you have no control? A very sensitive receiver is more prone to intermodulation and cross-modulation effects, and these may be more important.

At vhf the external noise levels are much lower and low-noise receivers are required for good weak-signal performance. Since it's relatively easy to build low-noise receivers for 50 MHz with modern semiconductors, there's no excuse for being limited by system noise figure on this band. A receiver with a 5 dB noise figure at 50 MHz, for example, when used with 100 feet (30.5m) of RG-8A/U transmission line, will provide a system noise factor of 4.38 (NF = 6.4 dB) at the antenna terminals. This is well below the external noise.

The 144-MHz example in **table 2** is hopelessly inadequate and represents at least 12 dB degradation over what can be obtained in practice. A receiver with a 1.5 dB noise figure on this band, when used with the 100 feet (30.5m) of RG-8A/U, will provide a system noise

factor of 2.54 (NF = 4.1 dB) at the antenna terminals which is still inadequate. A transmission line with 0.7 dB loss would bring receiver noise figure within acceptable limits, but it might be easier and less expensive to install a low-noise preamp at the antenna.

As pointed out earlier, the noise temperatures of antennas that are pointed into space for satellite communications (or EME) are much lower than for terrestial communications where the antenna is pointed at the horizon. This means that the receiver noise figures must be lower for maximum performance. Some parts of the sky are noisier than others due to the presence of noise sources, but the noise figure of the receiver should ideally be low enough that the system is galactic-noise limited. Following are the receiver noise figures to shoot for when designing receivers or converters for satellite communications on vhf.

frequency	galactic noise floor	noise figure
28 MHz	−125 dBm	8 dB
50 MHz	−130 dBm	5 dB
144 MHz	−139 dBm	1 dB
220 MHz	−140 dBm	0.7 dB
432 MHz	−141 dBm	0.2 dB

These figures are based on a 2.1-kHz bandwidth and assume a lossless transmission line. For more accurate calculations at low noise figures, the use of noise temperatures is recommended.

Although the topic of noise figure measurement is beyond the scope of this article, the simplest method of making the measurement is to compare receiver noise to the noise generated by a temperature-limited vacuum diode. This technique is easily applied in the home workshop and has been discussed many times in the amateur radio magazines.[14,15,16] Guentzler also described a noise measuring system which used a pilot lamp as a noise source.[17]

intermodulation distortion

Amplitude distortion occurs in an amplifier when the magnitude of the output signal is not exactly proportional to the input signal. Although amplifiers can be designed to be nearly perfectly linear over a portion of their operating range, every amplifier has nonlinearity which can cause distortion products or harmonics of the driving waveform. *Intermodulation distortion* or IMD is a type of amplitude distortion which occurs when a nonlinear amplifier is driven by more than one discrete frequency. Although the discussion here is limited to IMD in receivers, this is the same distortion which is used to define the linearity of ssb linear power amplifiers.

When an rf signal with varying amplitude is passed through a nonlinear device, many new products are generated. The frequency and amplitude of each component can be calculated mathematically since the nonlinear device can be represented by a power series expanded about the zero-signal operating point.[18] Although many products are generated, the ones of primary concern are the second and third. This can be demonstrated with a two-tone signal with outputs at f_1 and f_2, 14001 and 14003 kHz.

$$f_1 = 14001 \qquad f_2 = 14002$$
$$2f_1 = 28002 \qquad 2f_2 = 28004$$
$$3f_1 = 42003 \qquad 3f_2 = 42006$$

Although each of the harmonics fall well outside the passband of a receiver which is tuned to pass 14001 and 14002 kHz, the harmonics mix together to produce intermodulation products which do fall within the passband. The third-order products consist of

$$2f_1 - f_2 = 14000 \text{ kHz}$$
$$2f_2 - f_1 = 14003 \text{ kHz}$$

The fifth-order products consist of

$$3f_1 - 2f_2 = 13999 \text{ kHz}$$
$$3f_2 - 2f_1 = 14004 \text{ kHz}$$

The output spectrum is shown in **fig. 6**. Unless the nonlinearity of the amplifier

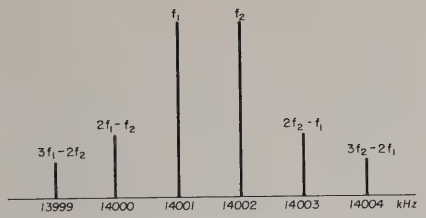


fig. 6. Third- and fifth-order intermodulation products generated by input signals at 14001 and 14002 kHz. In receiver stages fifth-order IMD is usually small enough to be neglected.

is particularly severe, fifth-order IMD is not usually a problem and can be ignored in receiver applications.

Although the IMD distortion products which are generated by two discrete frequencies are used here because they're easy to visualize, exactly the same sort of thing occurs with complex speech waveforms.

In a receiver rf amplifier or mixer stage IMD may be caused by two adjacent CW signals or by a ssb signal. Furthermore, in a mixer where the input must be wideband (such as the double-balanced mixer which is currently finding wide use), two input signals, (f_{R1} and f_{R2}) may mix with the local oscillator (f_L) to produce in-band, two-tone, third-order intermodulation products ($2f_{R1} - f_{R2}$) $\pm f_L$ and ($2f_{R2} - f_{R1}$) $\pm f_L$. Third-order intermodulation products also occur at ($2f_{R1} - f_{R2}$) and ($2f_{R2} - f_{R1}$). Two input frequencies at 14210 and 14230 kHz, for example, with a 5.2-MHz local oscillator (9 MHz i-f), will produce two-tone, third-order intermodulation products at 8990 and 9050 kHz (i-f passband) and 14190 and 14250 kHz (rf passband).

Third-order IMD is measured in the laboratory with a spectrum analyzer using the test setup in **fig. 7**. However, the concept of the third-order intercept is finding increased use to describe the IMD response of mixers, and can also be used to describe the linearity of amplifiers.* The third-order intercept is the theoretical point where the two-tone, third-order response is exactly equal to the two-tone input. Amplifiers and mixers are not operated at this level in practice, but the intercept point offers an internationally recognized figure of merit for comparison of devices, both active and passive.

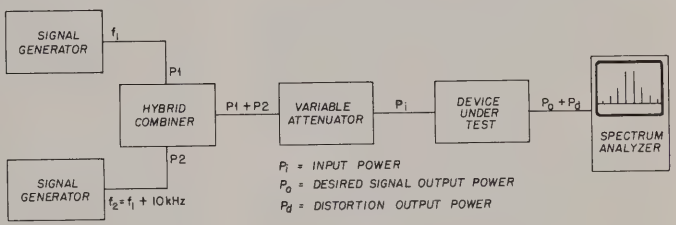


fig. 7. Block diagram of the test setup for evaluating the IMD performance of an amplifier or mixer with a spectrum analyzer.

In addition, the intercept point permits comparison of amplifiers and mixers where the intermodulation specifications are given at different two-tone levels. Once the intercept point is known, you can calculate the two-tone, third-order response at any input level by simply remembering that every 1-dB change in the two-tone input produces a 3-dB change in the third-order output. With this information it is possible to predict the maximum rf input level which is allowable.

With each 1 dB decrease in the f_R input level, for example, the third-order

*Class A amplifiers. Linear class AB or B amplifiers often exhibit two-tone, third-order intermodulation products which follow an S-shaped curve that both increases *and* decreases with additional input signal level so they cannot be compared by the intercept point method.

product is decreased an *additional* 2 dB. As shown in **fig. 8**, a high-level double-balanced mixer will suppress third-order products about 65 dB when both signals are at zero dBm (224 mV across 50 ohms) and 85 dB when both input signals are at −10 dBm (71 mV across 50 ohms). The third-order intercept point for these mixers is +27.5 dBm, relative to the *output*. Relative to the *input*, the intercept point is at +32.5 dBm. This is 17 dB higher than the intercept point for a low-level double-balanced mixer such as the Minilabs SR1A or Anzac MD108. The 3-dB compression point shown on the graph is a combination of both conversion compression and desensitization.

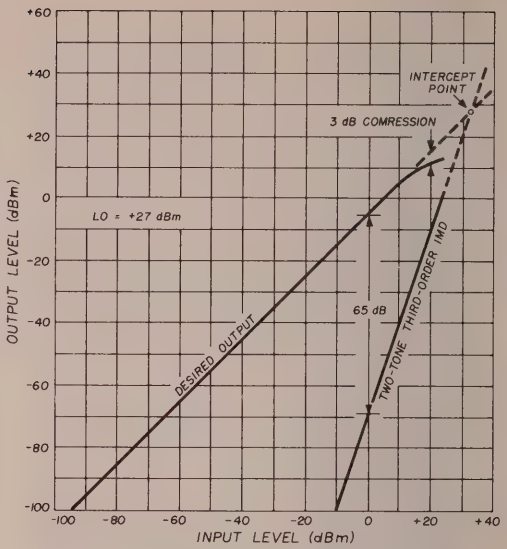


fig. 8. Third-order intercept point of high-level double-balanced mixer (SRA1H) at 50 MHz is +32.5 dBm relative to the input, 3-dB compression occurs at +20 dBm and third-order IMD is suppressed 65 dB when both input signals are 0 dBm (224 mV across 50 ohms).

intercept point

The third-order intercept point, IP, can be calculated from the relationship

$$IP = \tfrac{1}{2}(P_o - P_d) + P_i \qquad \textbf{(16)}$$

IP = third-order intercept, dBm
P_o = desired output, dBm
P_d = third-order distortion products, dBm
P_i = input power, dBm

Since the third-order IMD is defined as $(P_o - P_d)$, **eq. 16** can be rewritten as

$$IP = \tfrac{1}{2}IMD + P_i \qquad \textbf{(17)}$$

For the spectrum display shown in **fig. 9**, for example, the third-order intermodulation distortion products with two input signals of 4 mV (−35 dBm) are 80 dB down, and the third-order intercept is

$$IP = 0.5 \cdot 80 - 35 = +5\ dBm$$

Most amateurs don't have spectrum analyzers, but if good intermodulation distortion information is provided on the receiver data sheet, the intercept point can be easily calculated with **eq. 17** (in all too many cases, however, amateur receiver manufacturers ignore IMD completely, and when they do provide IMD data, it is incomplete). However, assume the specifications for an amateur-band receiver list −75 dB IMD for an input of 1 mV (−47 dBm). The third-order intercept is

$$IP = 0.5 \cdot 75 - 47 = -9.5\ dBm$$

Once the intercept point is known, the IMD performance at any input level can be found by rearranging **eq. 17**.

$$IMD = 2(IP - P_i) \qquad \textbf{(18)}$$

For an intercept point of −9.5 dBm, for example, the IMD at various input levels is shown below.

input signal	IMD
100 μV (−67 dBm)	115 dB
500 μV (−53 dBm)	87 dB
1000 μV (−47 dBm)	75 dB
5 mV (−33 dBm)	47 dB
10 mV (−27 dBm)	35 dB

Compare this with the state-of-the-art receiver front end described on page 26 of this issue which has −74 dB IMD at an input of 100 mV (−7 dBm). The

intercept point is at +30 dBm.

input signal	IMD
100 μV (−67 dBm)	194 dB
500 μV (−53 dBm)	166 dB
1000 μV (−47 dBm)	154 dB
5 mV (−33 dBm)	126 dB
10 mV (−27 dBm)	114 dB

The superiority of this receiver is obvious — for all but the strongest signals the IMD products are at or below the receiver noise level. Assuming a 10 dB noise figure and 2.1-kHz bandwidth, an input signal of −23.7 dBm (14.7 mV across 50 ohms) will produce IMD products just equal to the noise level. In the receiver with an intercept point at −9.5 dBm, however, the IMD products are already 3 dB greater than the noise with an input of −49 dBm (800 μV). This will be discussed further under the subject of dynamic range.

double-balanced mixers

Often a mixer data sheet does not specify the third-order intercept point, but a rule-of-thumb estimate can be easily made by examining the 1-dB compression point. As the rf input is increased, the i-f output should follow in a linear manner. However, after a certain point, the i-f output increases at a lower rate until the mixer output becomes fairly constant. The point at

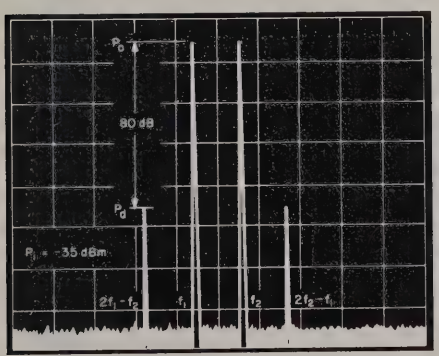


fig. 9. Spectrum analyzer display of communications receiver shows IMD is 80 dB down with a two-tone input of −35 dBm (4 mV across 50 ohms). Intercept point is at +5 dBm.

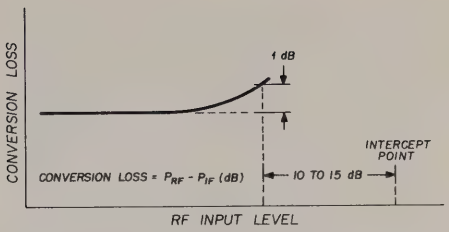


fig. 10. When working with double-balanced mixers, the third-order intercept point may be estimated by using the rule of thumb that the intercept point is 10 to 15 dB above the 1-dB compression point (see text).

which the i-f output deviates from the linear curve by 1 dB is called the 1-dB compression point. At this point the conversion loss is 1 dB greater than it was when the rf input was smaller.

The importance of the 1-dB compression point is its utility in comparing the dynamic range, maximum output and two-tone performance of various double-balanced mixers. As a rule of thumb, the third-order intercept point is approximately 10 to 15 dB higher than the 1-dB compression point[19] (about 15 dB at the low frequencies and 10 dB at higher frequencies). This is shown in **fig. 10**.

To properly use a double-balanced mixer, it is necessary to relate the two-tone input and third-order output levels to avoid generating excessive distortion which would compromise the final design. This is equally valid for amplifiers. Also important, but not as obvious, is the effect higher operating frequencies have on the double-balanced mixer's two-tone, third-order distortion characteristics. Performance is usually better at the lower frequencies and drops off as frequency is increased. For typical high-frequency double-balanced mixers with a maximum frequency specification of 500 MHz, performance starts to fall off somewhere between 50 and 100 MHz.

As discussed by DJ2LR on page 26, it is now possible, using a double-

balanced mixer in the front end, to build high-frequency communications receivers with a third-order intercept at +30 dBm. Using the rule of thumb that the 1-dB compression point is 15 dB below the intercept point, 1-dB compression occurs at an input of approximately +15 dBm or 1.25 volts across 50 ohms.

By comparison, the 1 dB compression point of many commercial amateur receivers is in the vicinity of −20 dBm (22 mV across 50 ohms) and some solid-state receivers with bipolar rf amplifiers go into compression at −40 dBm (2 mV across 50 ohms).

cross modulation

Another type of amplitude distortion which can occur in tuned amplifiers is *cross modulation*. This is related to IMD and is produced when the modulation from an undesired signal is partially transferred to a desired signal in the passband of the receiver. The 3-dB compression point in **fig. 8** describes the start of cross-modulation effects.

The cross-modulation effect is independent of the desired signal level and is proportional to the square of the undesired signal amplitude. Because of this relationship, an rf attenuator which lowers the signal level at the input to the receiver may provide a great improvement in cross-modulation performance. A 6 dB attenuator at the receiver input terminals, for example, will reduce cross modulation by 12 dB. If the desired signal is at least 6 dB above the level at which the receiver provides a satisfactory S/N ratio this results in a marked improvement in received signal quality.

Cross modulation is measured in the laboratory by setting one signal generator to deliver a CW output and another generator is set up for 30% amplitude modulation. The output of the a-m generator is increased until 1% modulation appears on the CW signal as measured with a spectrum analyzer. This represents a cross-modulation ratio of about 30 dB (cross-modulation level 30 dB below the reference level).

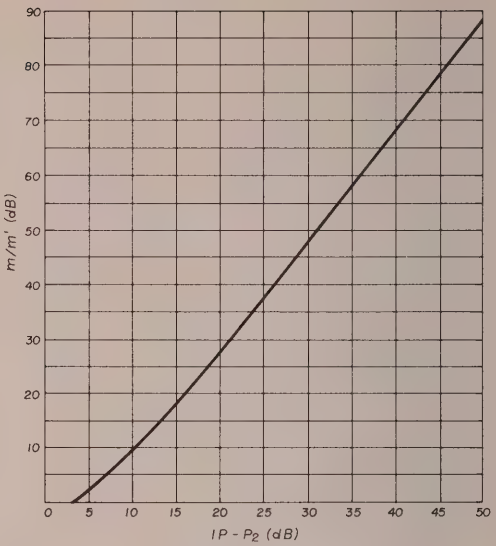


fig. 11. Graph of cross modulation vs the ratio of the intercept point to the interfering signal level in dB.

Cross modulation is related to the intercept point by the relationship

$$m/m' = (P_{ip}/4P_c) - \tfrac{1}{2} \quad \textbf{(19)}$$

m/m' = ratio of cross modulation transferred from a large signal to a smaller one

P_{ip} = intercept point power

P_c = interfering signal power

Cross modulation in dB is simply

$$m/m'(dB) = 20 \log (m/m') \quad \textbf{(20)}$$

In **fig. 11** cross modulation is plotted against the difference between the intercept point and the cross-modulating signal in dBm. Cross modulation of 30 dB, for example, corresponds to 21 dB difference between the intercept point and the signal producing the cross modulation. For a receiver with an intercept point at +30 dBm, a modulated input signal at a level of +9 dBm (630 mV

across 50 ohms) will produce 30 dB cross modulation. For a receiver with an intercept point at −9.5 dBm (the more usual case), an interfering signal level of −30.5 dBm (6.7 mV) will produce 30 dB cross modulation.

gain compression

When a receiver is tuned to a weak signal, a strong, adjacent signal may cause an apparent decrease in receiver gain. This is called *compression* or *desensitization* and occurs when the input voltage from the undesired signal is large enough to exceed the bias on an rf amplifier or mixer and drives the base (or grid) into conduction. This reduces gain, as shown by the compressed curve of **fig. 8**, and increases distortion. The rectified base (or grid) current can also be coupled back to the receiver's agc system which results in a further reduction in overall receiver gain.

Compression is measured by setting one signal generator to produce a CW signal and another generator, at a given frequency spacing, is adjusted to depress the desired signal a certain amount, usually 3 dB. Like cross modulation, however, a compression specification has little meaning if the frequency separation between the two signals is not specified.

Since both cross modulation and compression are caused by strong, undesired signals which are adjacent to the receiver passband, they can be controlled to a certain extent by the selectivity at the front end of the receiver. High receiver sensitivity, of course, is the antithesis of good cross-modulation and compression performance — this reinforces the argument for receiver noise figures on the order of 10 dB for the high-frequency range.

dynamic range

The front end of a receiver is subjected to a multiplicity of input signals which tend to intermodulate to produce a level of distortion products which is dependent on the magnitude of the incoming signals. Therefore, the upper end of a receiver's dynamic range is defined by the input signal level which produces third-order IMD products just equal to the receiver's noise level.

At the lower end dynamic range is limited ultimately by the noise figure of the receiver. Also important, however, is the way the receiver handles weak signals. Some linear rf amplifiers and mixers give good performance in the middle of their operating range but exhibit transfer curves that introduce considerable distortion at low signal levels. With proper design, however, this is not a problem, and the dynamic range of a receiver is usually defined as the *spurious-free dynamic range* where the maximum input signal is as defined above and the minimum input signal is at the noise floor of the receiver (**eq. 8**).

$$DR = 2/3\ (IP - N_o) \qquad \textbf{(21)}$$

where DR = spurious-free dynamic range, dB
IP = intercept point, dBm
N_o = receiver noise floor, dBm

The dynamic range of a receiver is important because it allows you to directly compare the strong-signal performance of one receiver against that of another. On today's crowded bands, and the high incidence of kilowatt transmitters and directive antennas, strong-signal performance is usually much more important than sensitivity. Although dynamic range can be used as a figure of merit, it's also useful to know the maximum input signal level, $P_{i(max)}$, which will produce third-order IMD products just equal to the noise level. This can be calculated from*

$$P_{i\,(max)} = 1/3(2IP + N_o) \qquad \textbf{(22)}$$

$P_{i\,(max)}$ = maximum input signal, dBm
IP = intercept point, dBm
N_o = receiver noise floor, dBm

* Derivation of **eq. 16** through **eq. 22** will be sent to interested readers upon receipt of a self-addressed, stamped envelope.

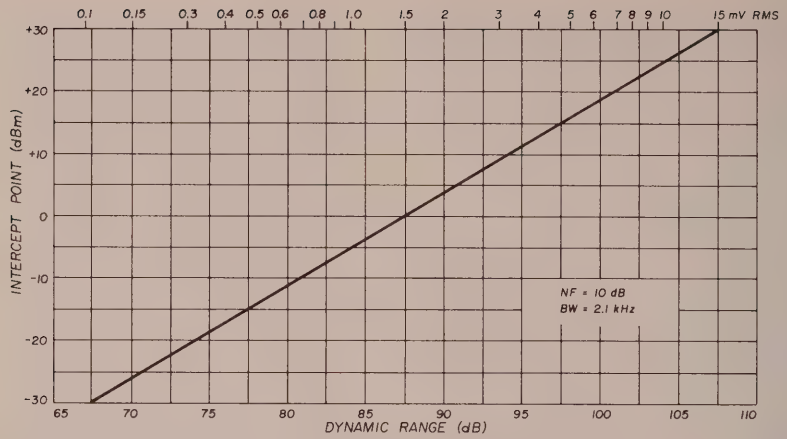


fig. 12. Spurious-free dynamic range vs intercept point (based on 2.1-kHz bandwidth, 10 dB noise figure). Maximum input signal (millivolts rms across 50 ohms) is shown at top.

For example, assuming a noise figure of 10 dB and 2.1-kHz bandwidth, the spurious-free dynamic range and maximum input signal of a receiver with an intercept point at +30 dBm are

$$DR = 2/3[30 - (-131)] = 107.3\ dB$$

$$P_{i(max)} = 1/3[2 \cdot 30 + (-131)]$$

$$= -23.7\ dBm$$

Thus, the IMD products will be well below the noise level for all input signals below −23.7 dBm (14000 μV or about S9+43 dB)! As a comparison, consider a receiver with a third-order intercept at −9.5 dBm (10 dB noise figure, 2.1-kHz bandwidth):

$$DR = 2/3[-9.5 - (-131)] = 81\ dB$$

$$P_{i(max)} = 1/3[2(-9.5) + (-131)]$$

$$= -50\ dBm$$

With this receiver the distortion products are equal to the noise level with an input signal of about 710 μV or S9+17 dB. This may represent adequate strong-signal performance if you live out in the country, but it's doubtful. If you live in an urban area, you'll have a lot of trouble digging weak signals out of the morass of IMD products which effectively raise the noise floor of the receiver.

A graph of intercept point vs dynamic range and maximum input signal is presented in **fig. 12** for a receiver with a 10 dB noise figure and 2.1-kHz bandwidth (typical for modern amateur communications receivers).

summary

Although modern amateur receivers are no longer performance limited by noisy vacuum tubes or poor selectivity, the published performance specifications have changed little since the 1940s and are still limited essentially to data on sensitivity and selectivity. Specifications on intermodulation distortion, cross modulation, desensitization and dynamic range, if they're mentioned at all, provide insufficient information for direct buyer comparison. Purchasing a new receiver under these conditions is a bit like buying a new car without knowing gas mileage or how many passengers it will carry.

The performance specifications for the high-frequency receiver shown in **table 3** leave no question as to receiver performance and are recommended as a

table 3. Specifications for a high-performance, high-frequency communications receiver provide a good format for amateur equipment manufacturers to follow. These specifications give complete reference and qualifying data and leave little question as to actual receiver performance.

Frequency range	500 kHz to 30 MHz
Tuning accuracy	±500 Hz relative to the frequency of the desired signal
Sensitivity	CW and ssb: 0.5 μV for 10 dB S/N ratio in a 2.4-kHz bandwidth (11 dB noise figure)
I-f selectivity	2.1 kHz at −6 dB, 4.2 kHz at −60 dB (2.0 shape factor)
Intermodulation products	**Out of band:** With two 20 mV signals separated and removed from the desired signal by not less than 25 kHz the third-order IMD products are not less than 90 dB below either of the interfering signals. Intercept point = + 24 dBm **In band:** Two in-band signals of 20 mV will produce third-order IMD products not greater than −50 dBm
Dynamic range	102 dB
Cross modulation	With a desired signal greater than 100 μV in a 2.4-kHz bandwidth, an unwanted signal, 30% modulated, removed not less than 25 kHz, must be greater than 175 mV to produce an output 30 dB below the output produced by the desired signal.
Compression	With a desired signal of 500 μv, an unwanted signal more than 25 kHz removed must be greater than 300 mV to reduce the output by 3 dB
Spurious response	**External** signals 25 kHz or more removed from the desired signal must be at least 85 dB above the level of the desired signal to produce an equivalent output **Internal** spurious signals are not more than 3 dB above the noise level measured in a 2.1-kHz bandwidth
AGC range	An increase in input of 110 dB above 1 μV will produce an output change of less than 6 dB

guide for receiver manufacturers to follow in the future. Well informed amateurs should demand nothing less.

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ham radio

high dynamic range receiver input stages

A new look at the input stages of high-frequency communications receivers including some circuits for improving the IMD performance over that of any present amateur equipment

Ulrich L. Rohde, DJ2LR, 425 Fairfield Avenue, Ridgewood, New Jersey 07450■

Various efforts have been made in the past to build high-frequency communications receivers with wide dynamic range to combat the effects of cross-modulation and blocking. With the high level of galactic and man-made noise which is present up to about 30 MHz, receiver designers generally agree that a receiver with a 10 dB noise figure is more than adequate for high-frequency communications. This is equivalent to a sensitivity of 0.2 μV for 10 dB signal-plus-noise to noise (50-ohm input impedance, 2-kHz bandwidth). In most cases, on the high-frequency bands, the man-made noise which is picked up by the antenna is greater than 0.07 μV, so the extraordinary sensitivity of a receiver with a noise figure lower than 10 dB is no great advantage. While this technique was initiated ten years ago by my company, Rohde & Schwarz, in Munich, amateur equipment designers have only recently started thinking along the same lines.

A noise figure of slightly less or equal to 10 dB can be readily achieved by using one of the new hot-carrier diode double-balanced mixers, such as the Minilabs SRA3H or SRA1H, which is

followed by a low-noise i-f amplifier ahead of a crystal filter. In practice this means that no rf amplifier stage is required. In vacuum-tube receivers one of the primary applications for the rf input stage was to provide agc but this can now be obtained (with negligible losses) with a PIN diode attenuator.[1,2]

The first amateur equipment to use this new design technique is the Atlas 180 and its later models. There are some indications that this circuit is derived from the SouthCom SC130 Man-Pack Transceiver* and the AN/PRC-104 of Hughes Aircraft, which both use the same technique. The main disadvantage of this rf input circuit is that it is practically impossible to suppress feedthrough energy from the oscillator to the antenna to 15 μV or less, a requirement of European regulatory agencies.

Some military and systems-oriented receivers still require noise figures of 6 dB or less because they are used with inefficient antennas (whip antennas with loading coils for marine, mobile or portable use, for example). Since it is difficult for the commercial communications equipment manufacturer to foresee whether the receiver will be used with a good antenna or a poor one, coupled with the requirement for low oscillator feedthrough, in many cases rf amplifier stages are still required.

Recently the German amateur journal, *cq-DL*, published an extensive test report on the Atlas 180[3] which showed fairly poor dynamic range with respect to what would be expected from a receiver with a double-balanced mixer input and high-power i-f stage before the crystal filter. As pointed out in a previous article,[4] to obtain the specified performance every mixer must see a purely resistive load at *both* the i-f and image frequency. An analysis of the SouthCom and Atlas circuit reveals a tuned circuit between the double-balanced mixer and the i-f stage. Since the mixer is not properly terminated at the image frequencies, this results in a loss of at least 15 dBm for the mixer's third-order intercept point, enhancing intermodulation distortion products.

table 1. Third-order intercept point of Minilabs SRA1H high-level double-balanced mixer with various terminations.

termination	intercept point
50-ohm resistance	30 dBm
Narrow-band resonant circuit	8 dBm
Heavily damped resonant circuit	17 dBm
Elliptical bandpass circuit	21 dBm
Amplifier (Z_i = 50 ohms ±10°, 1-80 MHz)	23 dBm
Power fet (Z_i = 50 ohms ±5°, 1-108 MHz)	30 dBm

*Atlas Radio is licensed by SouthCom International to use the circuits developed by SouthCom for military and commercial communications equipment. **Editor.**

mixer circuits

A recent survey published in Germany[5] showed the results of a series of tests regarding the creation of unwanted spurious in double-balanced mixers as a function of termination. All tests were run with a high-level Minilabs SRA1H mixer which requires a local-oscillator input of 23 dBm (200 milliwatts). The results of these tests are listed in **table 1**.

Although the Minilabs SRA1H is capable of providing a third-order intercept point of 30 dBm, **table 1** indicates

that you can lose as much as 22 dB of dynamic range because the mixer is not properly terminated. Therefore, to obtain maximum performance, it is essential to build an input stage which has the required input impedance from dc

be adjusted to set the dc bias so the input impedance is exactly 50 ohms.[5] This circuit has been designed for optimum dynamic range regardless of oscillator feedthrough and does not show any input selectivity which must, of course, be

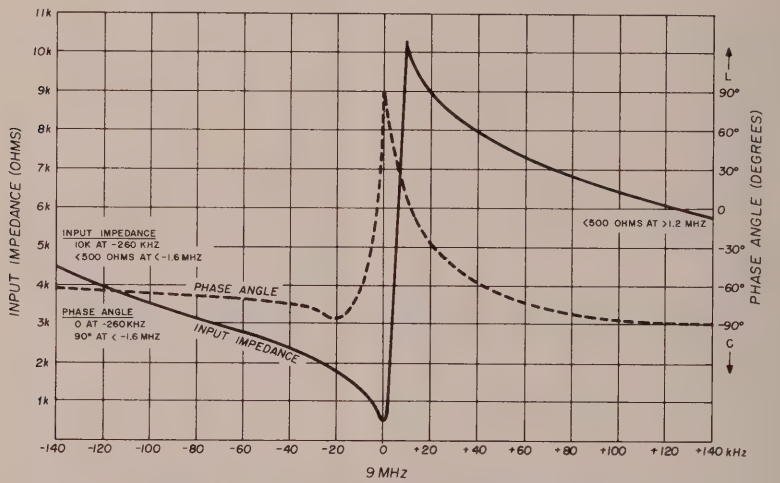


fig. 1. Input impedance and phase angle of the 9-MHz KVG XF9B crystal filter when terminated in 560 ohms in parallel with 33 pF.

up to more than 100 MHz. This stage must also have an intercept point of 25 dBm (the double-balanced mixer has about 5 dB loss so the mixer output is 25 dBm at 30 dBm input).

Extensive tests with various crystal filters have been conducted in the past and it was found that the crystal filter's input and output must be terminated beyond its normal frequency range of operation. This is because crystal filters, such as the KVG XF9B, at frequencies not too far from the center frequency, exhibit impedances between 5 and 10 kilohms (see **fig. 1**). These resonant effects would significantly reduce the dynamic range of the preamplifier if the filter was not properly terminated.

The best solution to the problem of a high dynamic range preamplifier is a high current field-effect transistor, type CP643 (Teladyne Crystalonics), in the circuit of **fig. 2** where the input resistors can

added. In most cases input selectivity can be provided by a 1.6-MHz highpass filter and a 31-MHz lowpass filter, so the losses there do not have to be considered.

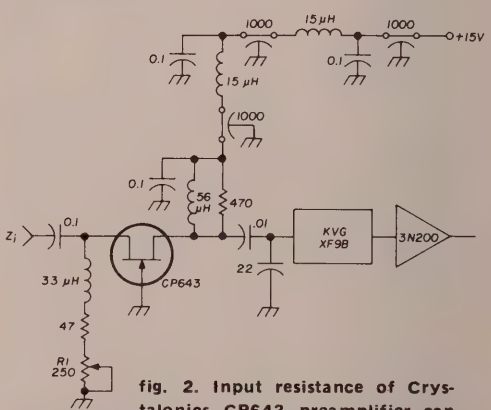


fig. 2. Input resistance of Crystalonics CP643 preamplifier can be adjusted to 50 ohms by proper setting of the 250-ohm potentiometer, R1. Current drain is 30 mA.

The complete rf input circuit, showing the CP643 high dynamic range preamplifier, double-balanced mixer, 2N5109 oscillator injection amplifier and 3N200 i-f amplifier is shown in **fig. 3**. The injection amplifier accepts about 200 mV input voltage from the local oscillator and amplifies it to the required level for the double-balanced mixer. The grounded-gate CP643 has 14 dB gain and is matched to the input of the KVG XF9B crystal filter. The 3N200 i-f amplifier has enough gain and agc action for most receiver designs.

push-pull rf amplifier with wide dynamic range

In many wideband, high dynamic range applications such as antenna distribution amplifiers, input rf amplifiers are required which combine extremely low distortion with low noise figure. In the past both voltage and current feedback have been used to counteract voltage and current distortion. The disadvantage of these circuits is that a stable input impedance can be achieved only over a relatively narrow bandwidth.[6]

Recent research has resulted in a new

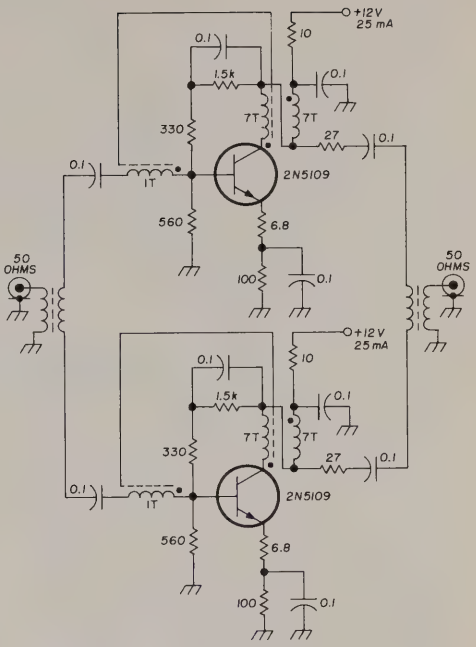


fig. 4. Vhf power transistors are used in push-pull circuit to obtain wide dynamic range shown in fig. 5. Transformers are trifilar wound on Indiana General F625-9-TC9 toroid cores.

wideband amplifier design[2,7] which has extremely low vswr at both input and output as well as low noise figure. The

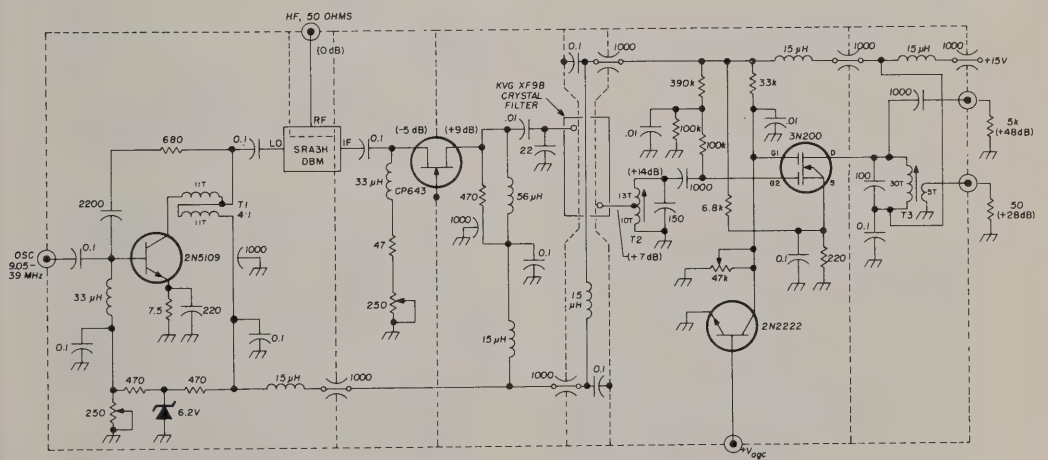


fig. 3. Double-balanced mixer circuit which has a third-order intercept point at +30 dBm. Oscillator requirement is −1 to +2 dBm (200 to 280 mV across 50 ohms). Agc range is greater than 50 dB. Levels shown in parenthesis are for zero dBm (224 mV) at input and zero agc voltage.

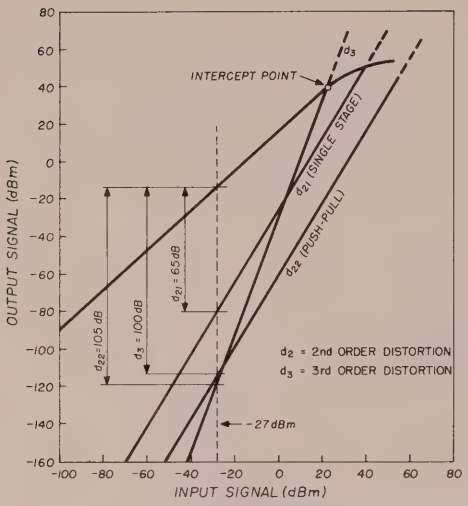


fig. 5. Performance of the push-pull rf amplifier shown in fig. 4. With an input of −27 dBm (two-tone signal, 20 mV each), gain is 12 dB, third-order distortion products are down 100 dB and second-order IMD is down 105 dB. Third-order intercept point occurs at an input of about 22 dBm.

second-order intermodulation products can be suppressed nearly 40 dB (over a single stage) by the push-pull arrangement shown in **fig. 4**. Two linear vhf power transistors are used in the circuit, and depending upon the large-signal handling requirements, either the 2N5109 (RCA) or BFR95 (Amperex) may be used. Both of these devices have an F_T of 1600 MHz.

This circuit provides about 11 dB gain and exhibits exceptional freedom from second- and third-order intermodulation products, as plotted in **fig. 5**. The third-order intercept point occurs at an input of about 22 dBm. Three types of feedback are used: *current* feedback through the unbypassed 6.8-ohm emitter resistor, *voltage* feedback through the unbypassed 330-ohm base-to-collector resistor, and *transformer* feedback through a third winding on the wideband transformer to stabilize the input and output impedance. A mathematical analysis of this circuit is presented for interested readers in the appendix.

table 2. Measured third-order intercept point of several commercial high-frequency receivers.

receiver	intercept point
Yaesu FT101	−21.5 dBm
Ten-Tec Argonaut	−19.5 dBm
Collins KWM2/S-line	−10.0 dBm
Signal 1 CX7	−5.0 dBm
Collins R390A	−4.5 dBm
Atlas 180/210	3.0 dBm
Collins 65S1	13.0 dBm
Racal RA1772	28.0 dBm
Martin rf front end (fig. 3)	30.0 dBm

summary

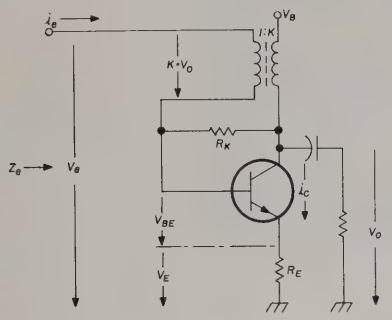
With very little effort the third-order intercept point of high-frequency receiver input stages can be increased far beyond the values of any commercial equipment now on the market. **Table 2** shows the third-order intercept point of several popular receivers presently being used by amateurs.

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1. Ulrich L. Rohde, "Eight Ways to Better Receiver Design," *Electronics*, February 20, 1975, page 7.
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3. Thomas Moliere, "Der Transceiver Atlas 180 - Testbericht," *cq-DL*, March, 1975, page 130.
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6. A. Hoenike, "Dimensionierungsfragen beim Entwurf klirrarmer Transistorverstaerker," *Nachrichtentechn. Z.*, May, 1966, page 287.
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appendix

Since voltage and current feedback may result in input and output impedances which may not suit the design requirements, a wideband toroidal transformer with a high-permeability core (such as Indiana General, F625-9-TC9) can be used to arbitrarily set these impedances. The schematic below



shows an amplifier using a transformer with voltage and current feedback. The input voltage, v_e, input current, i_e, and input impedance, Z_e, are given by the following equations

$$v_e = (k \cdot v_o) + v_{be} + v_e$$

$$i_e = (v_{be} \cdot Y_{11}) + v_{cb}/R_k$$

$$Z_e = \frac{(k \cdot v_o) + v_{be} + v_e}{(v_{be} \cdot Y_{11}) + v_{cb}/R_k}$$

So long as the operating frequency is well below f_T, emitter and collector current are the same. Therefore, the input impedance of the stage will be

$$Z_e = R_k \left(\frac{k + A}{1 - A}\right)$$

where $$A = \frac{R_E (R_k + R_L)}{(R_E + R_k) R_L}$$

and the required value for the voltage feedback resistor, R_k, is given by

$$R_k = Z_e \left(\frac{1 - A}{k + A}\right)$$

the amplification of the stage is given by

$$A = \frac{1}{k + \frac{R_E (1 + R_k/R_L)}{R_k + R_E}}$$

Since it is advantageous to have a 50-ohm input impedance, and the output impedance of the stage is approximately 150 ohms, the collector winding is split to build a 4:1 transformer. Under these conditions the input impedance, Z_e, output impedance, Z_o, and voltage gain, A, are given by

$$Z_e \approx (k \cdot R_k) + \frac{R_E}{2}$$

$$Z_o \approx \frac{R_E}{2k}$$

$$A = \frac{1}{k\left(1 + \frac{R_E}{2k \cdot Z_L}\right)}$$

The constant k, which determines the turns ratio between the base and collector coil (1:7:7 in **fig. 5**), can only be an integer. To obtain optimum performance in many cases, therefore, one of the values may have to be a compromise. In the circuit of **fig. 5** the output impedance was only 23 ohms so a 27 ohm resistor had to be placed in series to obtain the desired 50-ohm output impedance.

ham radio

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capacitance meter

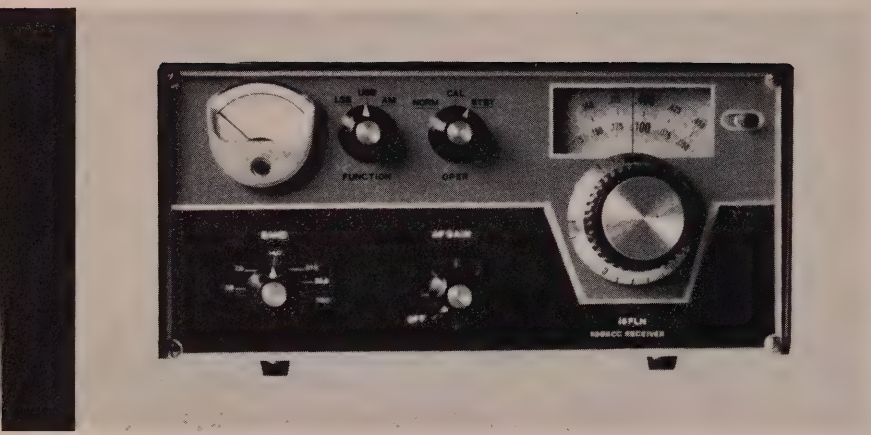
Dear HR:

I have receiver a number of queries regarding the programmable unijunction transistor used in the capacitance meter described in the April, 1975, issue of *ham radio*. The full part number of this device, which is manufactured by Texas Instruments, is A7T6028 (because of space limitations, the package is labeled AT6028). The 2N6027, 2N6028 and 2N6118 are similar.

Although I have not tried the Motorola HEP S9001 programmable unijunction in the circuit, I have letters from K6MYA and WA3IFQ who say they had excellent results with this device. Another possibility is the package of programmable unijunctions available from Radio Shack (part number 276-119).

The two series-connected 0.005 μF capacitors in the circuit may be replaced by a single 0.0025 μF capacitor. This apparently has caused some confusion.

Courtney Hall, WA5SNZ
Dallas, Texas



solid-state communications receiver

A five-band amateur-band receiver which features an active balanced mixer front end for improved intermodulation performance

Piero Moroni, I5TDJ, Cosseria 10, Florence, Italy 50129

Radio amateurs in Italy live in a special purgatory. The American ham magazines show us the beautiful equipment which is available, but customs duties and air shipment add about 100 per cent to the list price. This is probably one reason why we have a larger percentage of homebrew equipment here in Italy.

My first solid-state receiver[1] had been working fairly nicely, but I wanted something with a little better dynamic range, especially on 7 MHz as broadcast stations on this band are a little more bothersome here in Europe. A receiver with better cross modulation and blocking characteristics would help. The receiver described here seems to have achieved this objective, although I am an inveterate experimenter and will no doubt make a few changes. My friend Luciano, I5FLN, who duplicated my circuitry and packaged it very nicely, calls his copy the 5-Band DXCC reveiver (see photo above). He has worked over 150 countries on 7-MHz ssb. My re-

ceiver isn't as pretty — I was too impatient to hear the results!

The receiver uses a mixture of U.S. and Philips transistors which I accumulated, but I have suggested alternate types which are readily available in the States. Enterprising amateurs should have no trouble substituting equivalents from the myriad of types available. Performance characteristics of the receiver are listed in **table 1.**

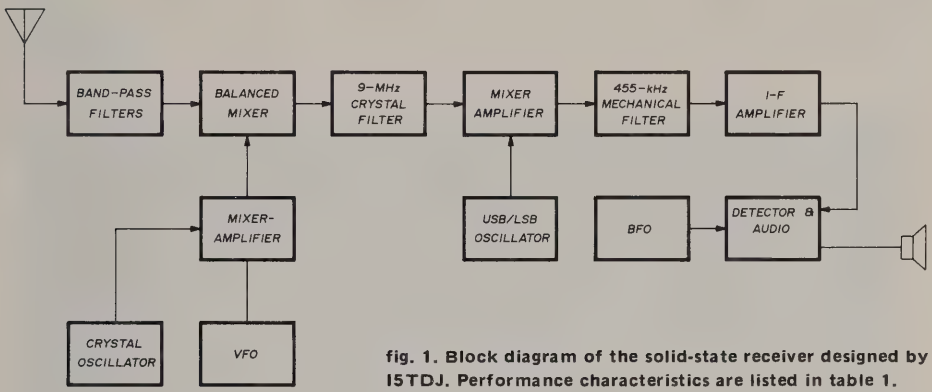


fig. 1. Block diagram of the solid-state receiver designed by I5TDJ. Performance characteristics are listed in table 1.

circuit description

A block diagram of the receiver is shown in **fig. 1.** The signal at the antenna first goes through bandpass filters, one for each band. Each filter consists of three capacitively-coupled tuned circuits which use high-Q toroids to reduce losses and suppress out-of-band signals.

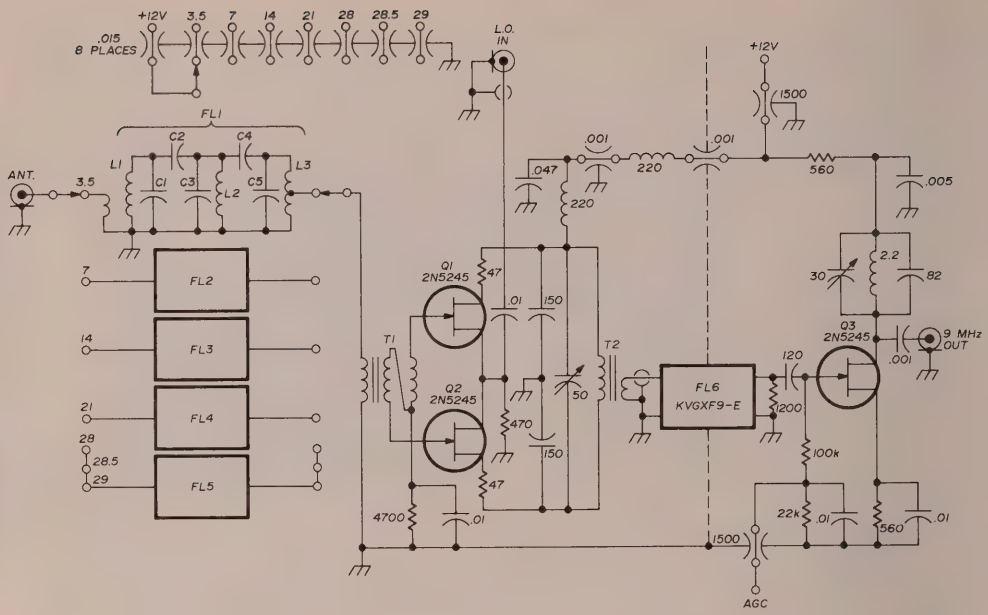
A balanced mixer following the filter insures that local oscillator noise is down some 30 dB. The vfo covers 5.0 to 5.5 MHz and effects the first conversion on 80 and 20 meters; for reception on 40, 15 and 10 meters the vfo is heterodyned with crystal oscillators for mixer injection.

The output of the balanced mixer is applied to a 9-MHz crystal filter, a KVG XF9E. This is an 8-pole filter with a 6-dB bandwidth of 2.4 kHz, a shape factor of 1.8 and stopband attenuation

This article was translated for *ham radio* by Josef Darmento, W4SXK. He reports that Piero is an engineer with a major electronic manufacturer in Florence, and that his ham shack is full of "auto constructed" (homebrew) equipment including some test instruments. Joe visited Piero in the fall of 1974 and had an opportunity to use the receiver described in this article. As Joe tells it, it was pure pleasure, but "how much was due to the receiver and how much was due to a different environment would be very difficult to evaluate."

table 1. Performance of the I5TDJ communications receiver.

Frequency coverage	80 through 10 meters
Sensitivity	0.25 μV for 10 dB signal-plus-noise-to noise ratio
AGC range	110 dB (3 dB increase in audio for antenna signal increasing from 1 μV to 300 mV)
Desensitization	70 mV (−13 dBm) signal ±100 kHz from desired 0.25 μV (−117 dBm) signal lowers signal-to-noise ratio from 10 dB to 3 dB
Intermodulation	With 5 mV (−33 dBm) signals ±50 kHz from desired 0.2 μV (−121 dBm) signal, distortion products are 88 dB below the interfering signals.



		L1	L2	L3	C1	C2	C3	C4	C5
FL1	3.5-4.0 MHz	14.2 μH. 59 turns no. 28 (0.03mm); 12-turn link	Same as L1, no link	Same as L1, tap 42 turns from ground	130	15	100	15	115
FL2	7.0-7.3 MHz	2.3 μH. 23 turns no. 24 (0.5mm); 5-turn link	Same as L1, no link	Same as L1, tape 17 turns from ground	200	8.2	200	8.2	200
FL3	14.0-14.5 MHz	0.97 μH. 16 turns no. 22 (0.6mm); 3-turn link	Same as L1, no link	Same as L1, tap 10 turns from ground	120	3.9	120	3.9	120
FL4	21.0-21.5 MHz	0.48 μH. 10 turns no. 22 (0.6mm); 2-turn link	Same as L1, no link	Same as L1, tap 7 turns from ground	120	3.3	120	3.3	120
FL5	28.0-29.7 MHz	0.48 μH. 10 turns no. 22 (0.6mm); 2-turn link	Same as L1, no link	Same as L1, tap 7 turns from ground	60	2.2	60	2.2	60

Note: All inductors for FL1 and FL2 are wound on Amidon T50-6 toroid cores; inductors for FL3 FL4 and FL5 are wound on Amidon T50-10 toroid cores.

T1 10 turns no. 32 (0.2mm), trifilar wound on Amidon T50-6 toroid core

T2 Primary is 22 turns no. 28 (0.3mm), evenly distributed on Amidon T50-6 toroid core; secondary is 7 turns no. 28 (0.3mm) over center of primary winding

fig. 2. Input filters, balanced mixer and 9-MHz amplifier. Filter capacitor values are in pF. Construction of this module is shown in fig. 3.

greater than 100 dB. A 9-MHz fet amplifier follows and a second fet mixer converts the 9-MHz signal to 455 kHz for application to a 2.7 kHz wide Collins mechanical filter. Crystal oscillators at 8545 and 9455 kHz provide sideband selection and are selected automatically with the bandswitch in accordance with current convention. A 455-kHz i-f amplifier, balanced diode detector, agc and an audio IC complete the picture. All circuitry derives power from an adjust-

able, regulated +12 volt power supply. Total current drain is about 375 mA.

construction

As indicated in the previous description, the receiver contains nothing new or startling. However, it is in the construction that careful attention to detail pays off. The receiver was built in modules, each module housed in a

mixer are on four small boards grouped around the bandswitch. The 9-MHz amplifier is further isolated in its own smaller box (see **fig. 3**). Bandpass filters F1 and F2 are on one board, F3 and F4 are on a second board, F5 is on a third and the balanced mixer is on a fourth. The objective here was to realize the maximum attenuation capability of the filters.

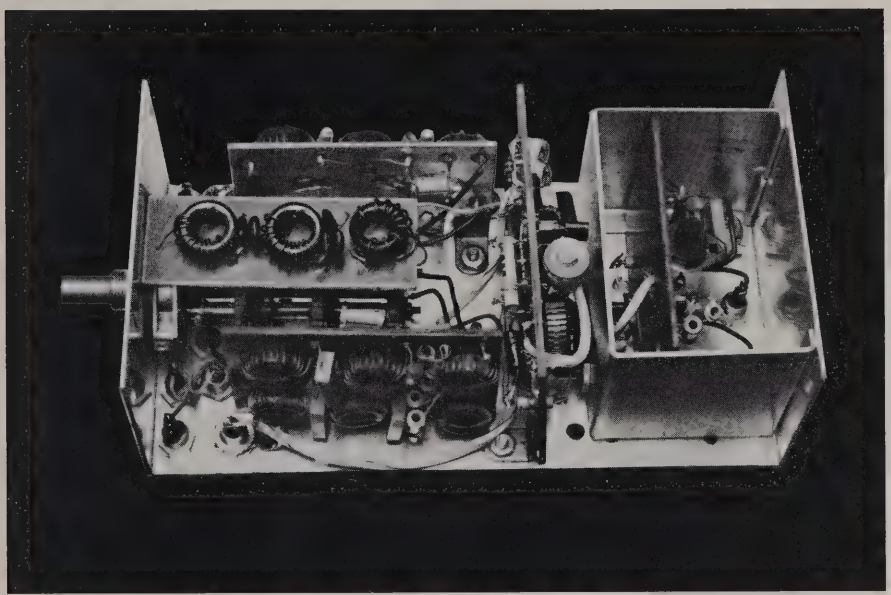


fig. 3. Mixer module showing the input filters (left) and 9-MHz amplifier (separate enclosure at right). The balanced mixer is on the vertical board in the center.

177x145x49mm (7x5.7x2 inch) high aluminum box, a commercially available size. Subdivisions within the boxes contain the functional sections, each on a piece of one-sided copper fiberglass circuit board. This was done to facilitate system modification. Modular construction also contributes to greater freedom from spurious products. All power leads enter the modules through 1500-pF feedthrough capacitors. Signal interconnections are by phone jacks.

mixer module

The input filters and the balanced

Transformer T1 is trifilar wound with a single primary and two secondaries in series. Transistors Q1 and Q2 should be selected for equal drain current with the gate at zero volts. The small box with the 9-MHz amplifier is completely shielded and isolated with a bottom and top cover of tinned copper sheet and is fastened with epoxy to the larger aluminum box. Only thus was I able to realize the maximum attenuation capabilities of the KVG filter; the filter itself is mounted externally. The gain of Q3 is agc controlled. Note that the leads carrying band-change informa-

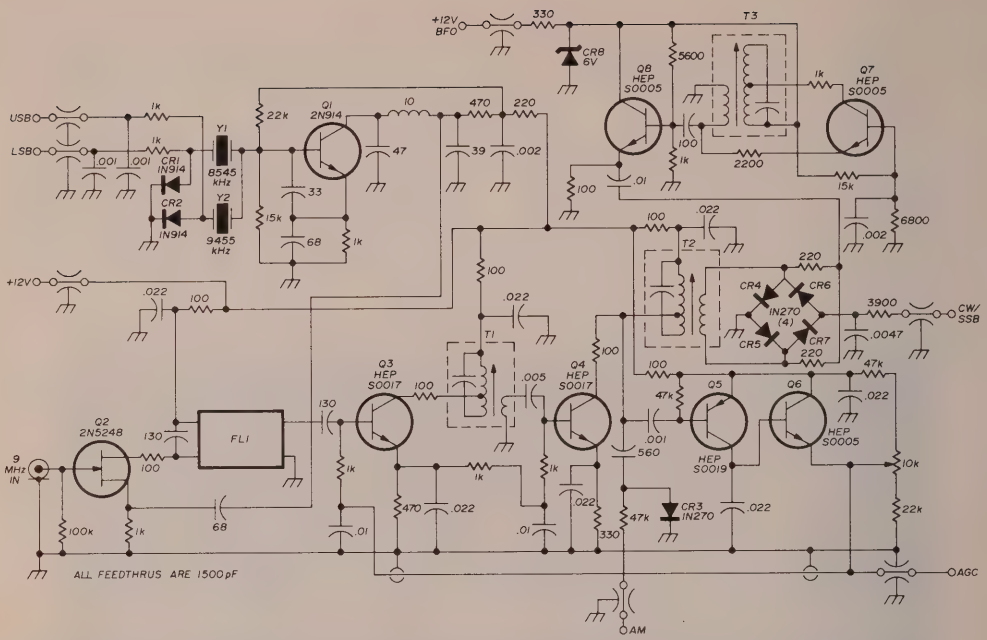


fig. 4. Circuit of the 455-kHz i-f module. The mechanical filter, FL1, is a Collins F455Z7. T1, T2 and T3 are miniature 455-kHz i-f transformers (see text). Module construction is shown in fig. 5.

tion enter the module through feedthrough capacitors.

455 kHz module

The 455-kHz amplifier is on a single board (see **figs. 4** and **5**). The Collins filter provides good isolation between input and output so no elaborate precautions are necessary. I mounted the transistors in sockets for easy substitution. The four diodes used in the product detector, CR4-CR7, are Phillips AAZ15 (1N270) selected for equal forward resistance. The BF175 transistors,

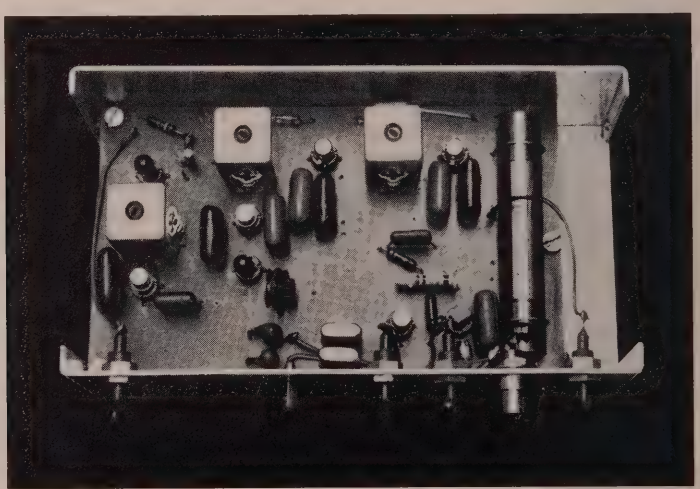


fig. 5. The 455-kHz i-f module. The Collins mechanical filter is at right.

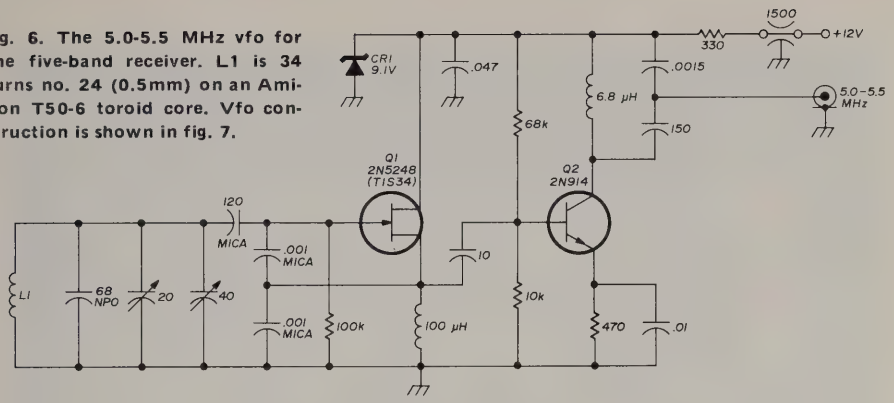


fig. 6. The 5.0-5.5 MHz vfo for the five-band receiver. L1 is 34 turns no. 24 (0.5mm) on an Amidon T50-6 toroid core. Vfo construction is shown in fig. 7.

Q3 and Q4, are rf types with gain which diminishes with increases in collector current. The Fairchild 2N4134 or HEP S0017 are good substitutes; these types may also be substituted for the BC209 and EC209. Note that transistor Q5, a BC154, is a pnp type; a HEP S0019 may be used instead.

Transformers T1, T2 and T3 are slug-tuned Siemens transformers — actually a bit large (25mm square) for this application. The J. W. Miller series 2031 i-f transformers only 0.5 inch square (12.5mm) might be more fitting if you have small fingers. The bfo may be crystal controlled if desired; its frequency, of course, will depend on the filter used in the receiver.

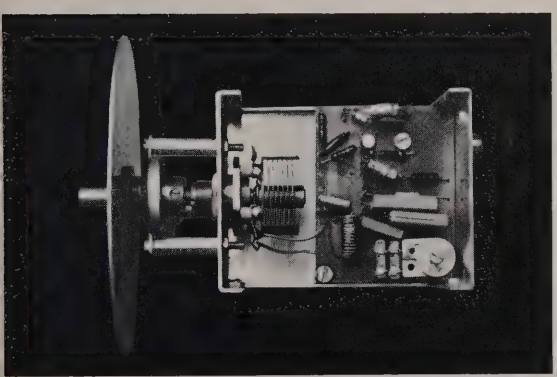


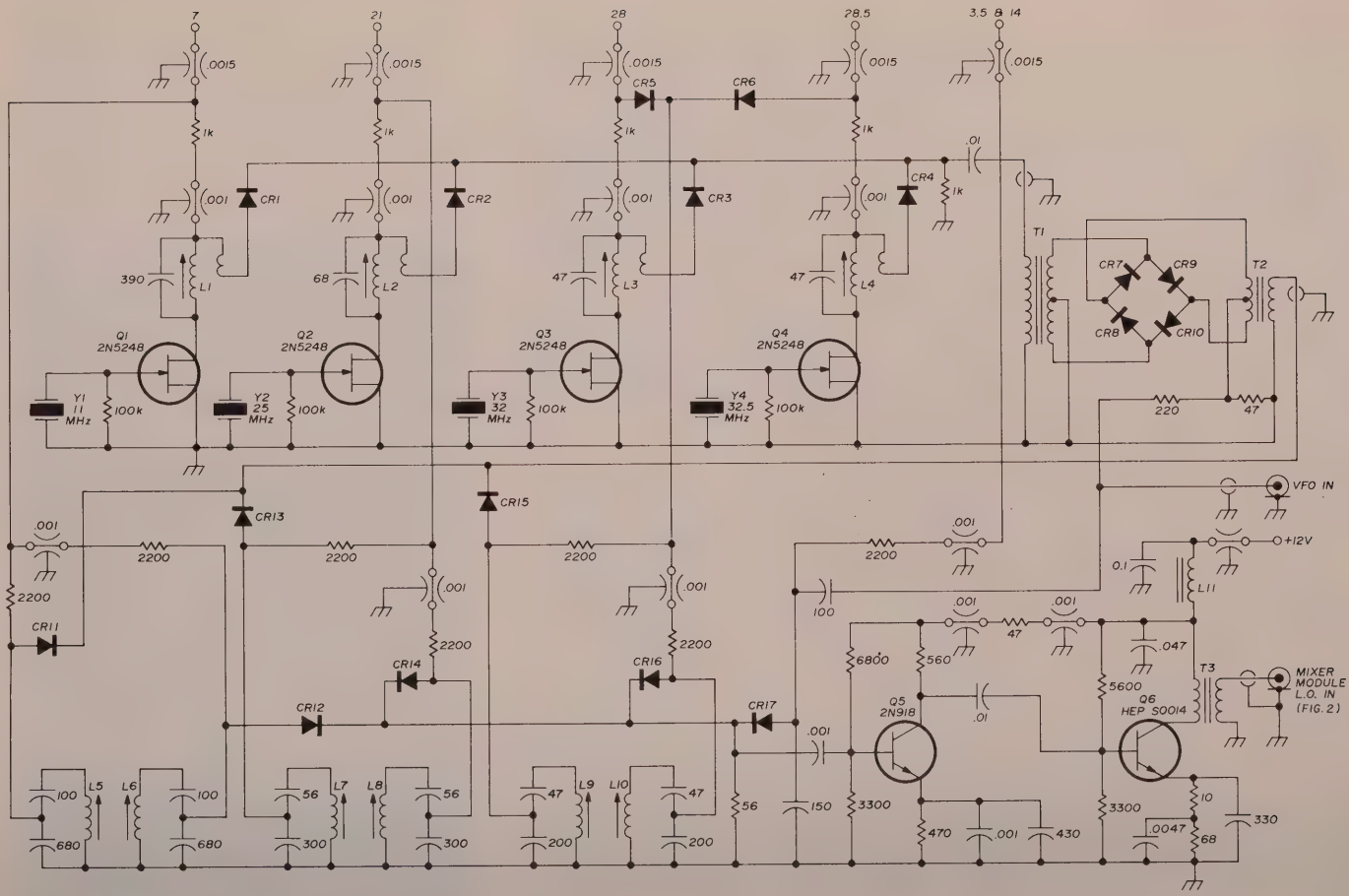
fig. 7. Construction of the 5.0-5.5 MHz vfo. A 5:1 planetary drive provides smooth tuning.

vfo module

The vfo is in a box which I strengthened by various methods to achieve mechanical rigidity (see **figs. 6** and **7**). The components are on a small fiberglass board except for the small 40 pF variable capacitor. The 68 pF NP0 affords temperature compensation and the 20 pF trimmer sets the band center. A 5:1 planetary drive gives nice easy tuning. A digital readout would have been even nicer, but would use a lot more power. Toroid L1 is installed edgewise on the board by its leads and a small dab of epoxy. I5FLN used a Drake vfo which makes an excellent substitution with less work.

vfo converter

The circuitry and construction of the vfo converter module are shown in **figs. 8** and **9**. The 5.0-5.0 MHz vfo signal at the vfo input goes directly to amplifiers Q5 and Q6 when the bandswitch is on 3.5 or 14 MHz. When applied to the balanced mixer the product is at 9 MHz. When the bandswitch is on 7, 21 or 28 MHz the vfo signal is mixed with the output of an fet crystal oscillator and filtered before being applied to Q5 and Q6. Two crystal oscillators at 28 and 28.5 MHz cover important segments of the 10-meter band. The filters suppress undesired products from the diode mixer, CR7-CR10. These diodes should


7
21
28
28.5
3.5 & 14
.0015
1k
.001
.01
CR1
CR2
CR3
CR4
CR5
CR6
390
68
47
L1
L2
L3
L4
Q1 2N5248
Q2 2N5248
Q3 2N5248
Q4 2N5248
Y1 11 MHz
Y2 25 MHz
Y3 32 MHz
Y4 32.5 MHz
100k
T1
T2
CR7
CR8
CR9
CR10
220
47
VFO IN
+12V
0.1
L11
CR11
CR12
CR13
CR14
CR15
CR16
CR17
2200
100
6800
560
5600
.047
T3
MIXER MODULE L.O. IN (FIG. 2)
Q5 2N918
Q6 HEP S0014
150
3300
470
430
10
330
.0047
68
L5
L6
L7
L8
L9
L10
100
680
56
300
47
200

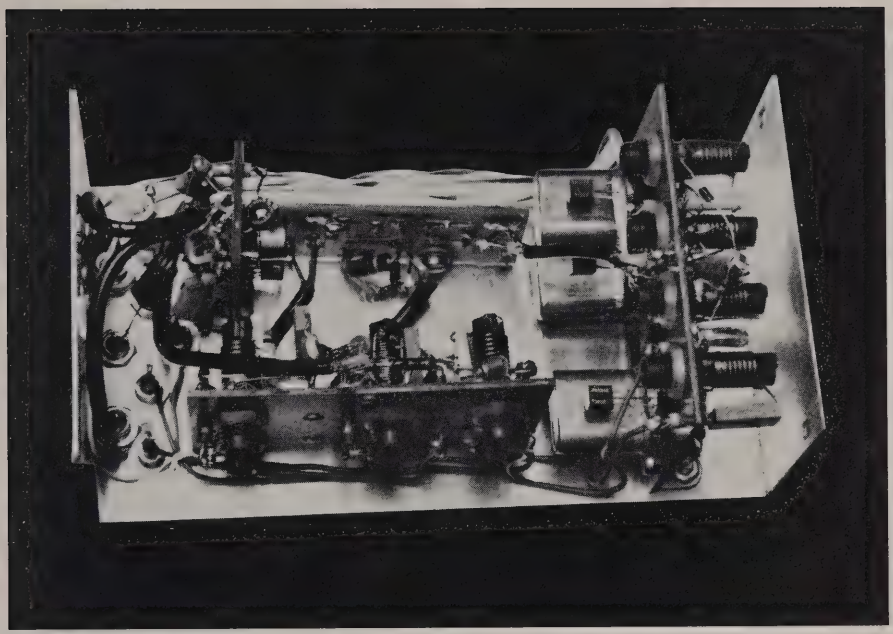


fig. 9. The vfo converter module. Filter pairs L5-L6, L7-L8 and L9-L10 are on the board in the foreground. Crystal oscillators are at the far right.

fig. 8. Vfo converter (left). All switching is accomplished with semiconductor diodes. Crystals are parallel resonant with a 32 pF load; Y2, Y3 and Y4 are third-overtone types. Construction of the module is shown in fig. 9.

CR1-CR6	1N914 or equivalent
CR7-CR10	Selected 1N270 diodes (see text)
CR11-CR17	1N914 or equivalent
L1-L4	0.6 μH (10 turns no. 22 (0.6mm) enamelled on 3/8" (9mm) diameter slug-tuned forms. Link is 3 turns no. 22 (0.6mm)
L5,L6	1.2 μH. 20 turns no. 28 (0.3mm) on 3/4" (9mm) diameter slug-tuned forms
L7-L10	0.6 μH (same as L1 - L4)
T1, T2	10 turns no. 32 (0.2mm), trifilar wound on Amidon T50-6 toroid core
T3	10 turns no. 32 (0.2mm), trifilar wound on Amidon T50-6 toroid core. Collector winding has two windings in series to give 2:1 ratio

be carefully selected for equal (±20 mV) voltage drop at varied values of current, say 0.75, 2, 10 and 20 mA. This is a must if you want oscillator attenuation of 30 to 40 dB.

The four fet oscillators are on one board and are energized by +12 volts from the bandswitch; diodes CR1-CR4 select the output. Drain coils L1, L2, L3 and L4 are wound on slug-tuned forms. The wideband trifilar transformers, T1 and T2, are on a second board with the diode mixer. The next board has the three filter pairs, L5-L6, L7-L8, and L9-L10. Each pair is inductively coupled by mounting the two coils about one diameter (9mm or 0.4 inch) apart. Input and output for the filters is via CR11 through CR17.

The fourth board has wideband amplifiers Q5 and Q6. The BFY63 transistor is a high-gain rf type which may be replaced by a HEP S0014. Since my

design called for amplification from 5 to 39 MHz I used as short leads as possible and ceramic capacitors. L11 is a small rf choke and consists of 20 turns or so on a 5mm (0.2 inch) ferrite rod. T3 is wideband and trifilar wound.

audio and power

Fig. 10 shows the rest of the receiver and **fig. 11** the interconnecting diagram.

alignment

After installation in their respective modules, I initially set most of my toroids with a grid dipper coupled with a one-turn loop, and a receiver for frequency accuracy. The job is simplified if you have a frequency counter. A slight loosening or tightening of the wire will bring the toroid to desired resonance. For a visual indication I used a vtvm

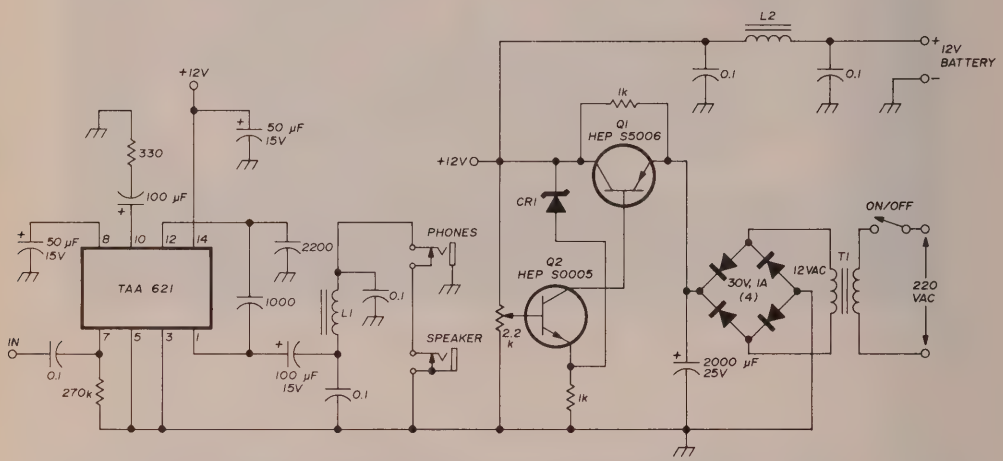


fig. 10. Audio amplifier and regulated power supply. A three-terminal 12-volt IC regulator such as the 7812 may be used, if desired. L1 and L2 are each 20 turns no. 18 (1mm) on a 5mm (0.2 inch) ferrite rod.

I did not bother enclosing the audio section or power supply. Since I am not a high-fidelity addict I used an audio IC which gives adequate volume for earphones or speaker. The TAA621 IC provides about 1 watt output with a 12-volt supply; an HEP C6093C or MC1454G may be substituted. L1 is a 5mm (0.2 inch) ferrite rod with 20 turns number-18 (1mm) wire; it and the two capacitors inhibit rectification of strong out-of-band rf signals (L2 in the power supply is the same).

I rewound an old transformer to give me about 12 volts and added a simple regulator. If you have a three-terminal voltage-regulator IC such as the MC7812, by all means use it.

with an rf probe. For the input filters I put the probe on the gate of Q1 and set the ±1 dB points of the filters as follows:

FL1	3.5 - 3.8 MHz
FL2	7.0 - 7.3 MHz
FL3	13.9 - 14.5 MHz
FL4	21.0 - 21.5 MHz
FL5	28.0 - 29.5 MHz

For inductances L1, L2, L3, L4 and L5 (**fig. 8**) I coupled the receiver to T1 and energized the respective oscillator. The probe on the hot side of T1 indicated about 0.5 volt for each oscillator as each coil was adjusted to resonance. Adjustment of the three filters is more easily done with vfo injection; alternately, a signal generator and rf voltmeter may be

used to set their -3 dB response as follows:

L5-L6	16.0 - 16.3 MHz
L7-L8	30.0 - 30.5 MHz
L9-L10	37.0 - 38.0 MHz

I used a calibrated receiver to set the center of the vfo excursion with the 20 pF trimmer; rf output was 0.6 volt. I connected the vfo to the converter input connector and temporarily installed

receiver S+N/N at 10 dB. I then introduced two more signals to the antenna, 5 mV each, at 28.75 and 28.8 MHz. The receiver output remained at 10 dB, i.e., distortion products were 88 dB below the two interfering signals.

For the blocking check I used one signal generator as before, 0.28 μV (-117 dBm) on 28.7 MHz, then introduced a second generator and adjusted

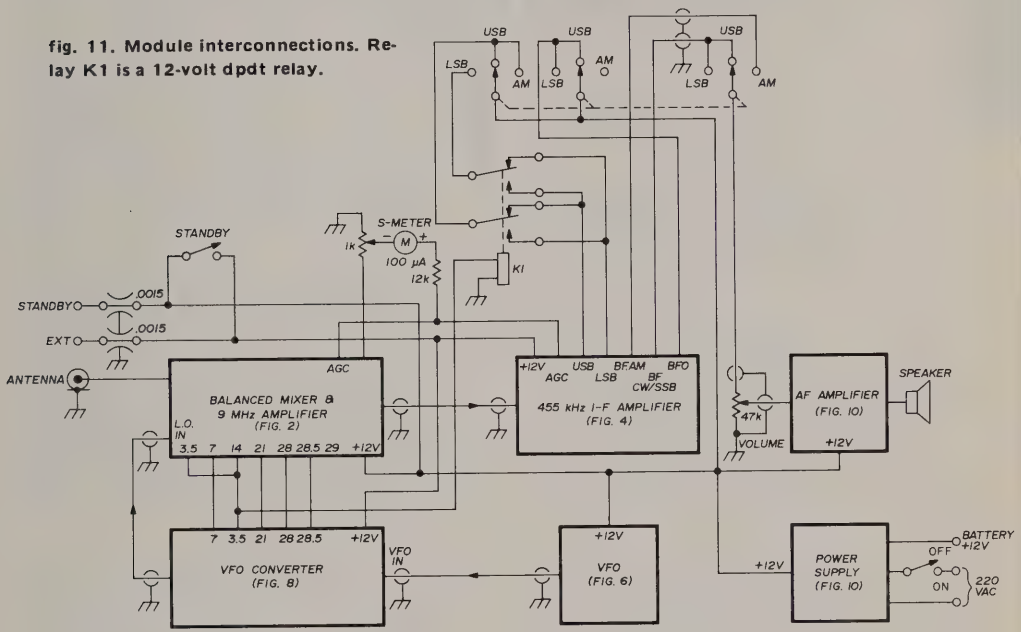


fig. 11. Module interconnections. Relay K1 is a 12-volt dpdt relay.

a 47-ohm, ½ watt resistor across the output jack. With the rf probe at this point and the bandswitch set on 80/40 I measured about 2 volts. The filters and drain inductances may be touched up so that the output stays at about 2 volts at any setting of the bandswitch. After interconnecting the other modules I made final adjustment following normal receiver alignment procedures.

It might be interesting at this point to describe how I measured intermodulation and blocking. For the intermodulation check I set one signal generator for 0.2 μV on 28.7 MHz and measured its frequency 100 kHz above, and 100 kHz below 28.7 MHz. In each case I advanced the amplitude of this second generator until the output S/N ratio deteriorated; when the amplitude was advanced to 70 mV the S/N ratio decreased to 3 dB.

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ham radio

low-cost 1296-MHz preamplifier

Low-noise, high performance 1296-MHz stripline preamplifier uses new Motorola MRF-901 microwave transistor

H. Paul Shuch, WA6UAM, Microcomm, 14908 Sandy Lane, San Jose, California 95124

A welcome side effect of the FCC's recent allocation of frequencies near 960-MHz to the Land Mobile Radio Service is the introduction by semiconductor manufacturers of low-cost, high quality active devices for the low end of the microwave spectrum. This technological revolution has proved to be a boon to amateur activity in the 1296-MHz band by bringing state-of-the-art components within the price range of microwave experimenters for the first time. This article presents construction details and performance data on a pair of preamplifiers for 1296 MHz which use the low-noise (under 2 dB), low-cost (under $10.00) Motorola MRF901 microwave transistor. Future developments will undoubtedly bring us a multitude of transistors offering superior performance and/or lower cost.

Numerous previous articles[1-5] have covered the design and construction of high quality preamplifiers for the 1296-MHz amateur band. These amplifiers used outstanding transistors from a number of different manufacturers. Designed primarily for military and aerospace applications, this generation of microwave transistors is characterized by high reliability over a wide temperature range, hermetic construction featuring a ceramic case with gold-plated leads, and (without exception) a cost far beyond the financial resources of the average experimenter.

The new transistors developed for the 960-MHz Land Mobile Band, with plastic cases and tinned-copper leads, are admittedly less rugged than their military predecessors; storage temperature is not above 150°C, breakdown voltages are lower and derating curves are steeper. These devices are, after all, intended for commercial service. Their reliability is, however, wholly adequate for an extended life in intermittent amateur service, and their electrical performance at 1296 MHz compares favorably with that of earlier devices costing an order of magnitude more.

design trends

Early solid-state uhf preamplifiers designed for amateur use achieved input and output impedance matching

through the use of pi networks composed of piston trimmer capacitors and slab inductors. This approach, typified by the designs of Katz[1] and Vilardi,[2,3] assured a proper impedance match regardless of the transistor characteristics because of the pi network's unique ability to "match anything to anything." Unfortunately, the pi network's very versatility made it difficult for the amateur who lacked the proper test equipment to know exactly when his amplifiers were tuned for optimum performance. Additionally, extensive tweaking of the input and output circuits made it possible to inadvertently adjust the amplifier into a condition of instability, with the resulting oscillations ultimately destroying the fragile transistor.

The microstripline designs of Donecker[4] and others changed all that. All matching transformers and reactances were etched onto a printed-circuit board, with no tuning adjustments whatever, so there was no need to optimize an amplifier on expensive test equipment. This approach made it practically impossible to destroy a transistor by inadvertently mismatching it. On the other hand, the lack of tuning adjustments made it impossible to compensate for minute differences between components or printed-circuit boards. And, since a board was computer designed to the parameters of a particular transistor, the experimenter had little opportunity

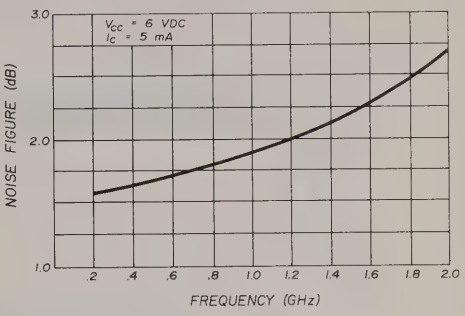


fig. 1. Noise figure vs operating frequency for the Motorola MRF-901 transistor.

to modify the design so he could use a different, more readily available device. In fact, some attempts to substitute transistors on the same circuit board led to a net degradation of the system noise figure to that of a simple diode mixer.

I attempted to rectify these limitations in my preamplifier designs by incorporating "tweaking" capacitors into a printed microstripline design.[5] By allowing one input and one output adjustment per stage, an amplifier is easily optimized, while lessening both the test equipment requirements and the likelihood of circuit instability. The same design approach is used in the preamplifiers presented here.

There is, however, one situation where the pi network matching technique still excels. For optimum system performance, it is necessary to match the input to the *first* preamplifier stage for optimum noise figure, *not* optimum gain. Optimum noise figure is achieved by deliberately and precisely mismatching the applicable input. Most rf transistor manufacturers publish curves or tables of complex impedances for pro-

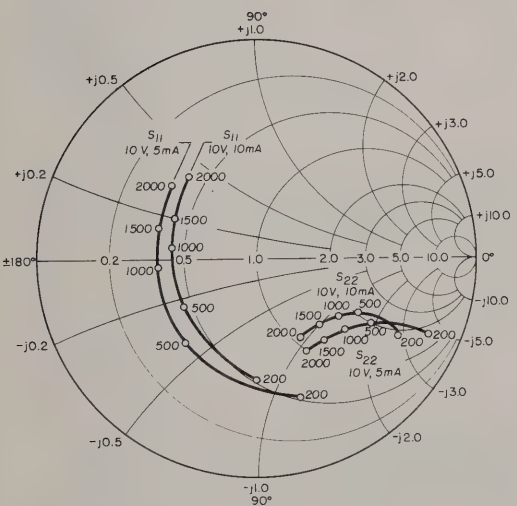
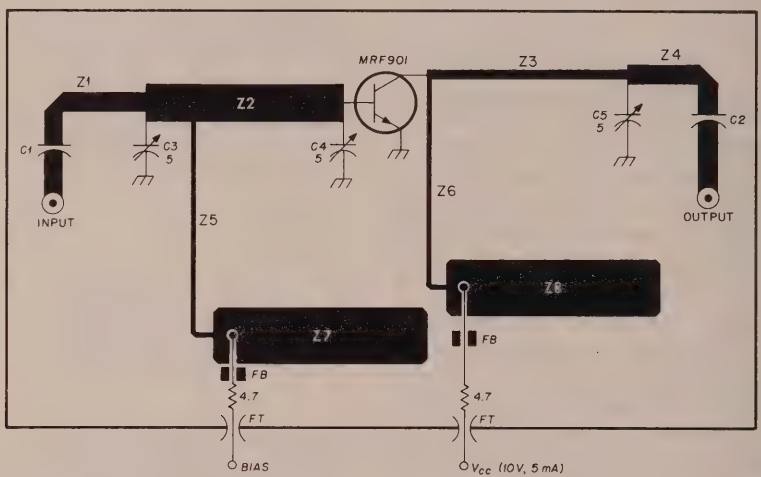


fig. 2. Input reflection coefficient, S_{11} and output reflection coefficient, S22, vs frequency for the Motorola MRF-901.

per input and output power match. Unfortunately, proper *mismatch* information for optimizing noise figure is seldom provided. Thus, it may be desirable to use a wide-range pi network at the input of the first preamplifier stage so that the source reflection coefficient at

figure optimized amplifiers around this transistor, and performance has consistently measured 2.8 dB (with the preamplifier looking into a high-quality double-balanced mixer). The 0.7 dB discrepancy is due, of course, to the familiar second-stage noise contribution,



C1,C2	Dc block. 50-150 pF chip capacitor (ATC 100B or equivalent)
C3,C4 C5	1-5 pF piston trimmer (Johannson JMC4642 or Triko 201-01M)
FB	Ferrite bead, slipped over resistor lead
FT	470-1000 pF feedthrough capacitor
Z1, Z4	50 ohm microstripline, 0.1" (2.5mm) wide, any convenient length
Z2	32.4 ohm, quarter-wavelength, microstripline, 0.20" (5mm) wide, 1.15" (29mm) long
Z3	74.8 ohm, quarter-wavelength microstripline, 0.045" (1mm) wide, 1.24" (31.5mm) long
Z5, Z6	Rf choke. 100 ohm, quarter-wavelength microstripline, 0.02" (0.5mm) wide, 1.25" (32mm) long
Z7, Z8	Rf short. 25 ohm, quarter-wavelength open-circuited microstripline, 0.30" (7.5mm) wide, 1.14" (29mm) long

fig. 3. Schematic of the 1296-MHz preamplifier. Capacitor C3 is used only with the noise-matched stage (see text). Quiescent collector current is 5 mA for the noise-matched stage, 10-mA for the power-gain stage. Full-sized printed-circuit layout is shown in fig. 5.

which noise figure is minimum can be empirically achieved.

transistor characteristics

Fig. 1 shows the manufacturer's claimed noise figure as a function of operating frequency for the MRF-901 transistor. At 1296 MHz the device is capable of delivering a noise figure below 2.1 dB. I have built several noise-

which was summarized quite well in the appendices to Reisert's recent article on 432 MHz preamplifiers.[6] Working through the textbook formula, it is found that the noise-figure optimized MRF-901 stage's intrinsic noise figure is on the order of 2.3 dB. Thus, the losses and mismatch errors inherent in this design still yield performance within 0.2 dB of optimum.

The specification sheets for the MRF-901 list input and output reflection coefficients (S_{11} and S_{22}) in polar form, tabulated for numerous combinations of frequency, power supply voltage and quiescent collector current. The two static conditions of greatest interest to the amateur are 10 volts at 5 mA (for a noise-matched first amplifier stage), and 10 volts at 10 mA (for a gain-optimized second preamplifier stage), as discussed in a previous article.[5] Reflection coefficients corresponding to these bias conditions were plotted on a Smith chart, then connected with a smooth curve as shown in **fig. 2.** The design parameters for the 1296 MHz preamplifiers were determined from interpolation of this Smith chart data.

design procedure

Two preamplifier stages are discussed here, one optimized for noise figure, the other for power gain. The gain-matched second stage was designed along lines analogous to that described in my previous articles. No noise-matched reflection coefficients were available from Motorola for the MRF-901, so a pi network was included in the input to the noise-figure matched stage, as discussed previously.

Fig. 3 is a functional schematic of the 1296-MHz preamplifier stages (the only differences between the stages optimized for noise figure and gain are the quiescent collector current [5 and 10 mA, respectively], and the use of an additional input matching capacitor, C3, in the noise-matched stage).

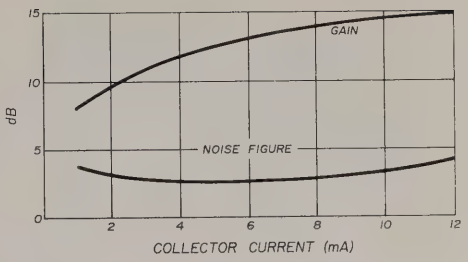


fig. 4. Typical power gain and noise figure of the 1296-MHz preamplifier as a function of collector current.

The system noise figures of several gain-matched 1296-MHz preamplifiers I've built with MRF-901 transistors all measured between 3.2 and 3.5 dB, while yielding 12 to 14 dB of power gain. Thus, for all but the most critical applications, the use of a noise-figure matched first preamplifier stage may not be necessary (see **fig. 4**).

construction

Fig. 5 is a full-sized printed-circuit layout for either preamplifier stage. The only visible difference between the gain-

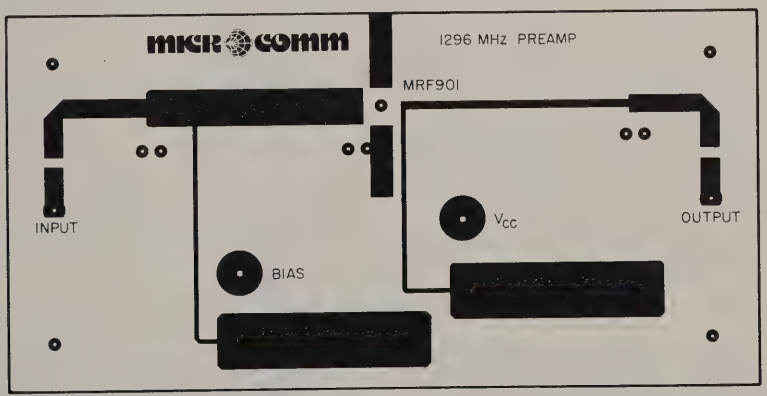


fig. 5. Full-sized printed-circuit board for the 1296-MHz preamplifiers using MRF-901 transistors. Component layout is shown in fig. 3.

matched and noise-matched stages is the incorporation of a variable capacitor, C3, in the noise-matched stage. Note, however, that the quiescent collector current differs between the two stages. More on this later.

The amplifiers are built on 1/16 inch (1.5mm) G-10 fiberglass-epoxy circuit board, double-clad with 1 ounce copper.

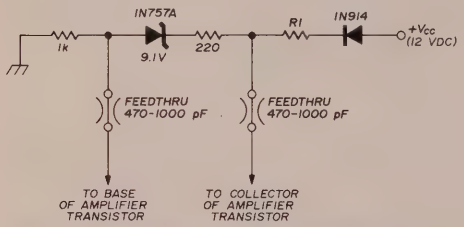


fig. 6. Zener bias circuit, developed by W1JAA, for use with transistor preamplifier stages. For 10 mA collector current, R1 = 120 ohms; for 5 mA collector current, R1 = 240 ohms.

I still receive occasional letters from readers who question my repeated use of glass-epoxy material at 1296 MHz. I will concede that the 0.2 dB excess noise figure above optimum which I have mentioned previously *may* be due to losses in the substrate. However, I feel that 0.2 dB is a small price to pay for the convenience and ready availability of this low-cost printed-circuit material.

Construction of these amplifiers is substantially the same as that of my previously published circuits. If you are unfamiliar with the fabrication of microstripline amplifiers, you are urged to refer to reference 5 for specific construction hints, as well as suggestions for tuneup and operation. That material is not repeated here.

bias circuit

The zener bias circuit introduced by Reisert[6] is not only simpler but also electrically superior to the active bias scheme I used in some of my earlier designs. Reisert's circuit is presented in **fig. 6** along with component values for the two required bias conditions. I refer the interested reader to his very fine article for a complete description of the operation of this bias circuit.

One appealing feature of this biasing scheme is that the power supply voltage can be varied upward or downward, as required, to optimize the stage for gain or noise figure. In this manner the performance of the preamplifier can be readily tailored to the requirements of the system in which it is installed.

conclusion

As the commercial microwave communications industry expands, sophisticated amplifiers have come within the reach of the average amateur experimenter. For those who prefer not to build their own equipment, low-cost 1296-MHz preamplifier modules are now available. A commercial version of the gain-matched stage, featuring a guaranteed maximum noise figure of 3.2 dB is currently available for under $40 from Microcomm.*

*For full specifications send a self-addressed, stamped envelope to Microcomm, 14908 Sandy Lane, San Jose, California 95124.

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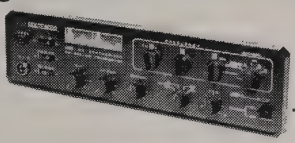
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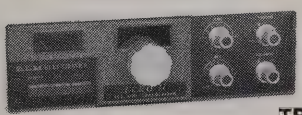
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low noise figure 28-30 MHz preamplifier for satellite reception

Inexpensive dual-gate mosfet circuit provides 20 dB gain and typical 1.0 dB noise figure

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Most modern communications receivers for the 28-30 MHz range are designed with dynamic range as the primary goal. Low noise figures generally require more gain ahead of the mixer and hence decrease the dynamic range. Furthermore, noise figures below 10 dB are seldom usable at these frequencies because the local ambient noise usually is the limiting factor for weak-signal detection.

Satellite reception and vhf/uhf converters usually require lower noise figures than the typical communications receiver will provide. Therefore, for these applications it's necessary to install a low-noise preamplifier ahead of the receiver. A suitable preamp which is inexpensive and easy to build is described in this article. It has proven to be a workhorse for many Oscar 6 and 7 operators and has also been used as a low-noise i-f preamp for use with vhf and uhf converters.

Many of the requirements for low-noise preamplifier design were discussed in an earlier article[1] and will not be reiterated here. Field-effect transistors were the prime candidates for this application because they are inexpensive, readily available, and yield low intermodulation distortion.

circuit discussion

The three preamplifier configurations shown in **fig. 1** were evaluated to determine which was the most suitable for low-noise operation at 28-30 MHz. Models of each were built and tested. The grounded-source circuit (**fig. 1A**) was discarded since it required neutralization for unconditional stability. Loading the drain circuit helped, but gain had to be reduced considerably before complete stability was obtained.

Next, the grounded-gate circuit (**fig. 1B**) was tried. Although it was stable, gain was lower than expected and the noise figure was slightly higher than the grounded-source configuration. The cascode circuit (**fig. 1C**) seemed to fill the bill; low noise, high gain and stability

were easily obtained using two single jfets *or* a dual-gate mosfet.

Noise figures of 1.5 to 2.0 dB were relatively easy to obtain with this circuit. The optimum source (input) impedance is 2 to 4000 ohms. After calculating the unloaded and loaded Q of such circuits (**fig. 2A**), I concluded that the noise figure was limited by input circuit losses, not device limitations. I tried the L-network input circuit (**fig. 2B**) and was pleasantly surprised by the results. The circuit losses were definitely lower. An rf choke which is parallel resonant at the operating frequency was used in the final configuration shown in **fig. 2C**. This method of matching is one

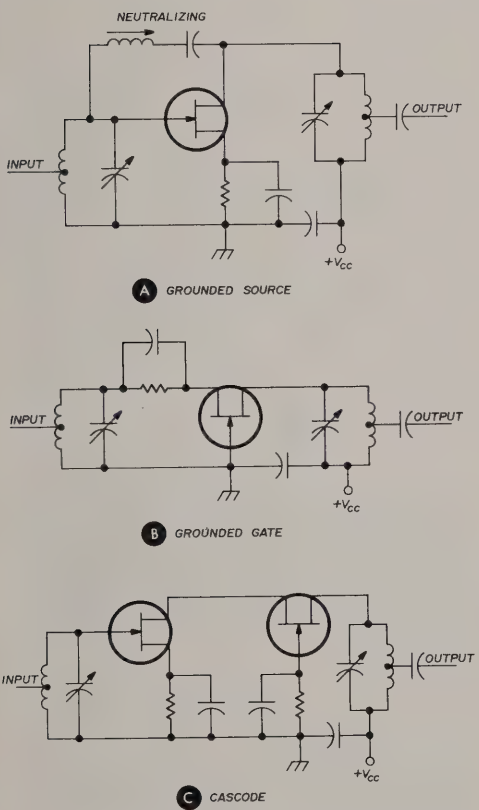
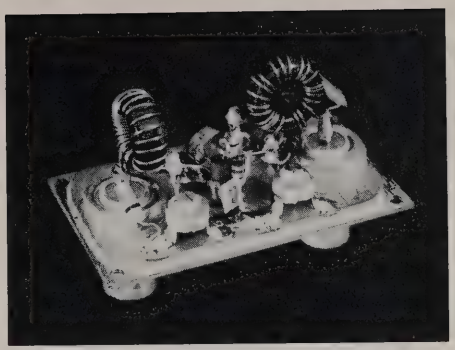


fig. 1. Fet preamplifier circuits which were evaluated for use as a low-noise 28-30 MHz preamplifier. Cascode circuit at (C) provides low noise, high gain and good stability.



Construction of the low-noise 28-30 MHz preamplifier. Note that the toroidal input and output transformers are installed at right angles to one another.

of the secrets to the low noise figure of this preamplifier.

device selection

The original circuit (**fig. 1A**) used a jfet. The Siliconix E300 and popular 2N4416 transistors worked equally well. These devices also worked well in the cascode circuit (**fig. 1C**). The Fairchild FT0601, a dual-gate mosfet, worked as well in the cascode circuit as the E300 types. Typical noise figures of 1.25 to 1.75 dB were easily obtained but the real surprise came when the inexpensive, diode-protected, dual-gate RCA 40673 mosfet was tried in the circuit. It consistently yielded lower noise figures than any of the other devices I tested: typical measured noise figures were less than 1.0 dB.

circuit description

The original preamplifier circuit had limited publication in March, 1972, and was revised a year later.[2] A complete schematic of the latest version is shown in **fig. 3**. The input network is as described earlier. RFC1 is parallel resonant at 28-30 MHz; values of 15 and 30 microhenries should be usable with little degradation. Inductors L1 and L2 are low-loss toroidal coils wound on Amidon cores, are small, and provide high

Q. Hence there is very little mutual coupling between the input and output circuits. Capacitor C1 allows optimization of noise figure and usually can simply be peaked for maximum gain. Capacitor C2 tunes the output circuit to the desired operating frequency. The circuit values shown in the schematic will permit operation from about 25 to 35 MHz without any component changes.

Attention is called to the use of a ferrite bead on the gate 2 lead to the mosfet, Q1. Careful testing revealed that most dual-gate mosfets tend to oscillate at uhf, typically at 750 to 900 MHz. This can be readily observed on a uhf spectrum analyzer. The reason for this is easily understood when you consider that there is a small, finite inductance present between gate 2 of the semiconductor dice and the outside of the transistor package. All attempts to bypass this lead were unsuccessful. The ferrite bead, however, did the trick since it essentially puts a lossy element on the lead and suppresses the uhf oscillation.

This same oscillation effect has also been observed on other dual-gate mosfet preamplifiers operating at frequencies as high as 500 MHz. In all cases a ferrite bead on the gate lead solved the problem.

Resistor R2 can be added to the circuit, if desired, to increase bandwidth with a small sacrifice in overall gain. Diode CR1 is an *idiot* diode which prevents damage to the mosfet if the power supply polarity is accidently reversed.

construction

Construction of the 28-30 MHz low-noise preamp is very straight forward and is similar to that used for the low-noise uhf preamp described in reference 1. A small section of double-sided printed-circuit board is attached to the cover of a small aluminum box (such as the Pomona 2417). BNC connectors are used for the input and output. The use of toroidal inductors eliminates the need for any shield if the input and output inductors are located at opposite ends of the box and are oriented at right angles to one another. Further construction details are shown in the photograph.

Although the RCA 40673 is a diode-protected mosfet, care should be exercised when handling the device to prevent any possible damage from static electricity. The best procedure is to first grasp the device by the case and then pick up the circuit board. With this technique the potentials are equal and the mosfet leads can be safely soldered into the circuit.

Any low-loss trimmer capacitors may be substituted for C1 and C2. The miniature JFD units I chose were based on

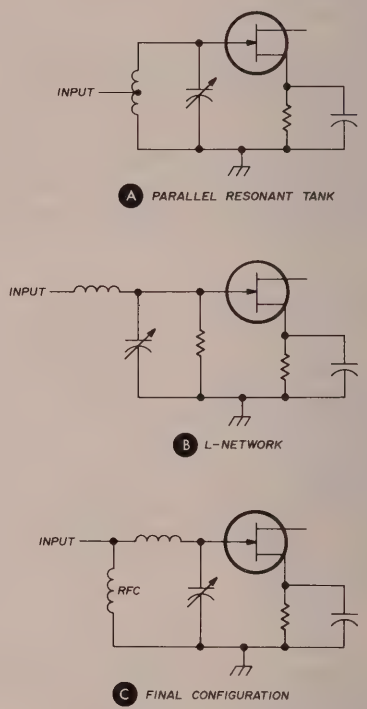


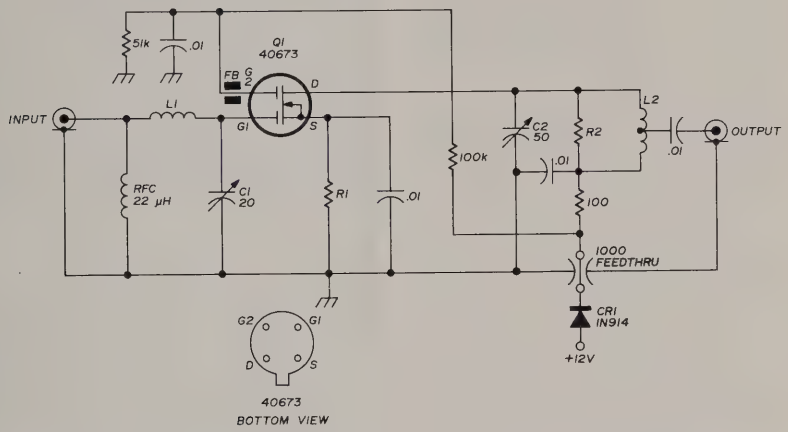
fig. 2. Input circuit configurations. Losses in the tapped, parallel-resonant circuit at (A) measureably increased the noise figure. L-network at (B) resulted in lower circuit loss. The circuit at (C), which uses a parallel-resonant rf choke, provided best performance and was used in the final preamplifier design shown in fig. 3.

the small size needed to package the circuit in the Pomona 2417 enclosure.

operation and test

For final checkout the input to the preamp should be terminated with a 50-ohm resistor, a noise generator or a 28-30 MHz antenna. The output can be connected to a receiver or i-f amplifier,

This preamplifier has been widely duplicated, both for reception of Oscar 6 and 7, and as a low-noise i-f preamp between vhf/uhf converters and a communications receiver. Gain is typically 20 to 25 dB and the highest noise figure I've seen is 1.25 dB. Most preamps have a maximum noise figure of 1.0 dB and some have measured as low as 0.7 dB.



C1 20 pF trimmer (JFD DVJ300 or equivalent ceramic trimmer)

C2 50 pF trimmer (JFD DVJ305 or equivalent ceramic trimmer)

FB Ferrite bead (56-590/65/3B or equivalent)

L1 25 turns no. 24 (0.5mm) on Amidon T50-10 toroid core

L2 22 turns no. 24 (0.5mm) on Amidon T50-10 toroid core, tapped 7 turns from cold end

R1 150 ohms typical (see text)

R2 2000 ohms typical (see text)

fig. 3. Low-noise 28-30 MHz preamplifier which provides 20 to 25 dB gain and typical noise figure of 1.0 dB uses a dual-gate mosfet in the cascode circuit.

The current drawn from the 12-volt power supply should be checked for proper operation: 3.0 to 7.0 mA is fine. If the current is too low or too high the value of R1 should be raised or lowered accordingly.

The only adjustments are to peak the input and output capacitors (C1 and C2, respectively) for maximum output. If a noise generator is available, C1 can be adjusted for lowest noise figure. However, as noted previously, in most cases good low-noise performance can be obtained simply by adjusting C1 for maximum gain.

The extremely low cost of this unit makes it a real bargain. Once you have one in operation, you'll wonder how you ever did without it.

Special thanks go to all those who have duplicated this preamp and encouraged me to write this article. I hope its use will improve your station's performance.

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bfo multiplexer

for a multimode detector

Cmos logic oscillator deck and phase-locked loop are combined for a multimode detection system

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For several years I've been experimenting with solid-state circuits to replace various receiver subsystems at my station. Most of the stages of my venerable HRO-50T have been rejuvenated in this fashion. Installation of this multimode detection system was a major step toward providing the performance and operating conveniences demanded of today's best receivers.

In the bfo multiplexer a single IC functions as an upper-sideband crystal oscillator, lower-sideband crystal oscillator, tunable bfo for CW, or as a limiter of the i-f signal for fm or synchronous a-m reception. The desired oscillator (or the limiter) is gated on by grounding its digital contol line. Multimode reception results when the multiplexed output of the oscillators and limiter is applied to a product detector. Fm signals can also be demodulated if the NE561 is used as a combination product detector and phase-locked loop discriminator.

Construction details are provided for a 455-kHz multimode detection system using the NE561. The IC module is packaged inside the original bfo coil's shield can for ease of installation. It is most economical, considering the features it provides, since surplus FT-241A crystals may be used and the ICs are not expensive.

cmos logic oscillators

Crystal-controlled oscillators using linearly biased cmos logic elements are described in RCA's applications literature.[1,2] In the basic circuit, shown in **fig. 1**, R1 biases the logic element as an inverting linear amplifier. The pi-type feedback network provides 180 degrees phase shift at the crystals's parallel resonant frequency to produce oscillation. The crystal sees a load capacitance equal to the series equivalent of C_s and C_t.

With regards to stability, the cmos oscillator circuit shown here is competitive with good discrete component de-

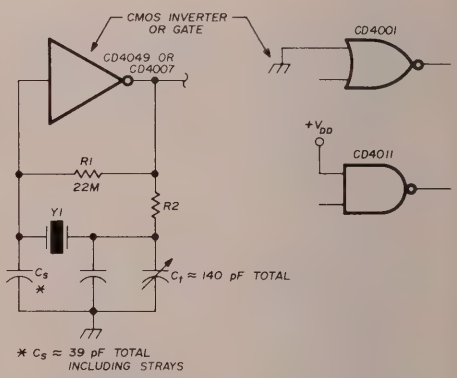


fig. 1. Basic crystal oscillator circuit using a cmos inverter or gate. Crystal Y1 is 10 MHz or less.

signs. The prototype of a 3.3 MHz oscillator that I am using in an industrial application, for example, is built around a high quality crystal and drifted less than 20 Hz between 0°C and 70°C without any temperature compensation. Power supply sensitivity has been reported as 3.5 ppm for a 25 per cent change in supply voltage.[1] Power drain, although proportional to frequency, is extremely low and a wide range of supply voltages can be accommodated.

Resistor R2 in **fig. 1** is the only component whose design value must change with frequency. At high frequencies it must be reduced or eliminated since the attenuation and phase shift it introduces increases with frequency.

multiplexer circuit

The idea which gave birth to the bfo multiplexer was that of obtaining four amplifier-oscillators, each with an on-off control, from a single quad 2-input NOR gate package and combining their outputs in a resistive summing network. The resulting circuit is shown in **fig. 2**. The supply voltage, V_{DD} should be between 3 volts and 15 volts. Regulation is necessary only to avoid exceeding the 15-volt upper limit. With a 9-volt power supply and the component values shown, a 770 mV p-p square wave output was observed. If you must use different summing network component values, use about 1000 ohms per volt of the supply in series with each output of the CD4001 IC to avoid overloading it.

Transformer T1 in the tunable oscillator circuit is a Radio Shack transistor oscillator coil. However, any high-Q coil-capacitor combination which resonates at the i-f will do. The stability of this circuit, particularly warm up drift, was noticeably better than that of the original circuit.

The input from the i-f amplifier should be at as high a level as possible for good limiting. This consideration should present no problem since the high input impedance of the CD4001

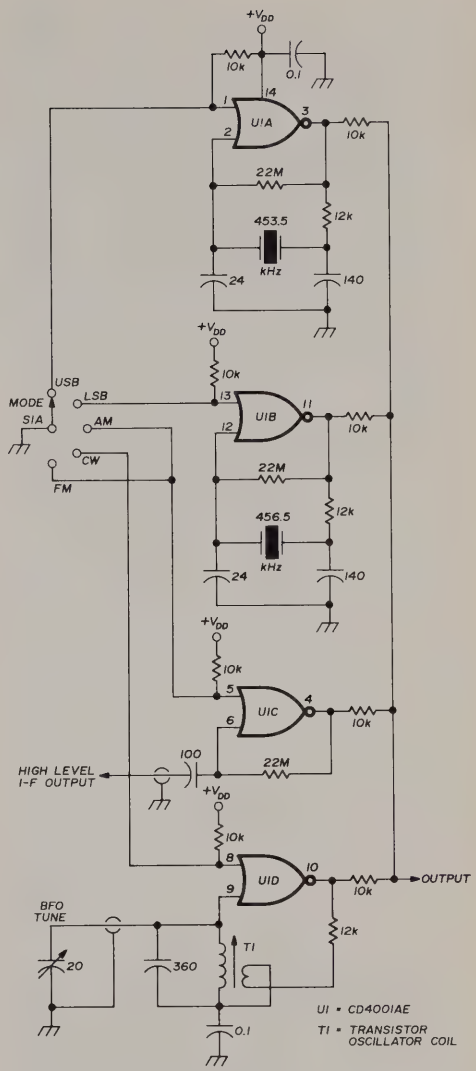


fig. 2. Schematic diagram of the bfo multiplexer. The supply voltage, V_{DD}, should be between +3 and +15 volts.

will not load a high-Q tuned circuit. Sensitivity for full output is about 300 mV p-p. The input level is adequate if a rounded square wave is obtained at the output with an a-m signal tuned in.

detector circuit

The bfo multiplexer provides a-m detection capability for a receiver equip-

ped with only a product detector while simplifying mode switching. In this application it replaces the original bfo. I first used the bfo multiplexer with a MC1496G product detector.[3] The results were excellent. Gain and dynamic range of this device exceed that of the NE561. A further advantage was that both the detector and bfo could use the same supply voltage. On the other hand, the gain and dynamic range were not needed and fm detection capability was desired.

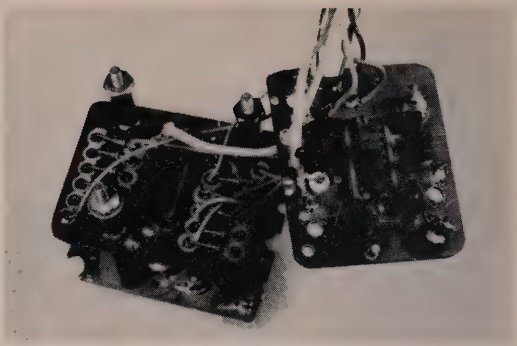
When used with the NE561, a 90 degree phase shift network is required in series with the output of the bfo multiplexer as shown in **fig. 3**. During a-m reception the network compensates for the fact that the NE561 locks up in quadrature with the signal at its phase detector's input.

The i-f input level to the NE561 should be held to about 100 mV rms for minimum distortion. If this is done the audio output level will be at least half that for narrow-band fm, about the same for ssb and CW, and at least twice for a-m if both sidebands are passed by the i-f filters. Note that the fm audio output level is proportional to the percentage deviation and cannot be increased by increasing the signal level. The two 0.004 μF capacitors limit the audio bandwidth to about 4 kHz.

The vco output of the NE561 is a 0.6 volt p-p square wave at the a-m carrier or bfo frequency. It appears to be ideal for use as an input to a digital dial adapter for accurate received frequency measurement.

construction

The multimode detection system was built on three 2 - 1/2 X 1-3/4 inch (57x45mm) copper-clad epoxy boards. The boards were stacked with phenolic posts, then slid inside the original bfo coil's shield can. The top board was bolted to the top of the shield can. Since the boards fit snugly inside the can, the complete assembly meets mechanical stability requirements. There is still room in the shield can for a fourth board containing a cable connector, if you are so inclined.



Closeup view of the detector board, right, and the first multiplexer.

The crystals and associated components are mounted in an octal socket on the top board. The oscillator coil and its associated components are mounted nearby. The top of the can is cut out to allow access to the crystals and the coil's adjustment screw without disassembly. Space could have been found for two trimmer capacitors for the crystal oscillators, had this been considered necessary.

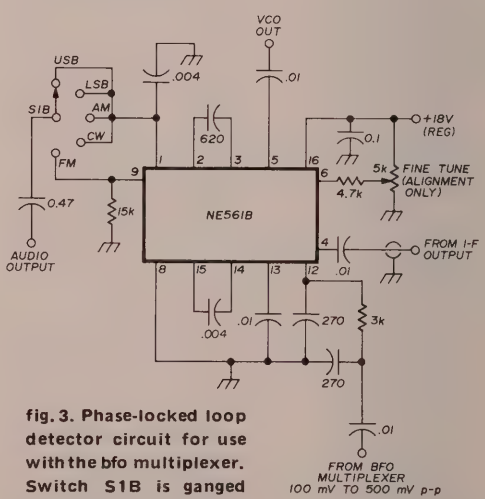


fig. 3. Phase-locked loop detector circuit for use with the bfo multiplexer. Switch S1B is ganged with the bfo multiplexer mode switch, S1A, in fig. 2.

The CD4001AE is mounted in a 14-pin DIP socket on the middle board. The two rows of resistors which flank the socket in the photograph are the pull-up resistors, summing resistors, and 22M feedback resistors. The pads were made with a Vector pad cutting tool.

The layout of the detector board can also be seen in the photograph. The fine-tuning control is adjacent to pin 16 of the IC. With the bfo multiplexer installed in the receiver, it can be reached through a hole drilled in the chassis. Alignment consists of setting the free-running frequency of the NE561 (bfo multiplexer off or disconnected) to within 200 Hz of 455 kHz using this control.



The completed multimode detector is housed in the original bfo shield can of the author's receiver.

operation

This detector system may be operated just like any other product detector-bfo combination. Most of the time you will be completely oblivious to the fact that there is a phase-locked oscillator in the system. However, when the limited i-f signal is applied to the NE561, several peculiarities may become evident. If conditions are just right, a coded CW signal may be demodulated in the a-m position as a series of chirps as the vco falls into and out of lock on each dit or dah. Another anomoly results when an interfering heterodyne is coincident with one of the voice sidebands of an a-m signal. Depending on their relative strengths, the loop may lock up on the interference, rather than the a-m carrier.

Excessive lock range can also be a problem. If strong signals hold the vco until you have tuned well past them, seemingly putting hysteresis in the tuning mechanism, a circuit modification may be in order. I found that a 22k resistor between pin 7 of the NE561 and its power supply reduced the lock range to ±6 kHz. A less drastic solution is to switch to the CW oscillator those few times it is necessary to bring the vco back on frequency.

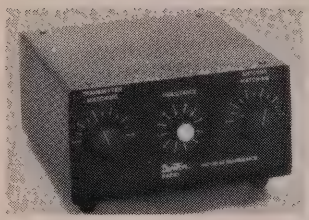
These phenomenon are simply quirks and are not reasons for not taking advantage of phase-locked loops for multimode detection applications. Other types of detectors have other problems under the same conditions.

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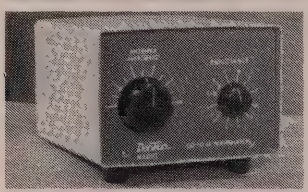
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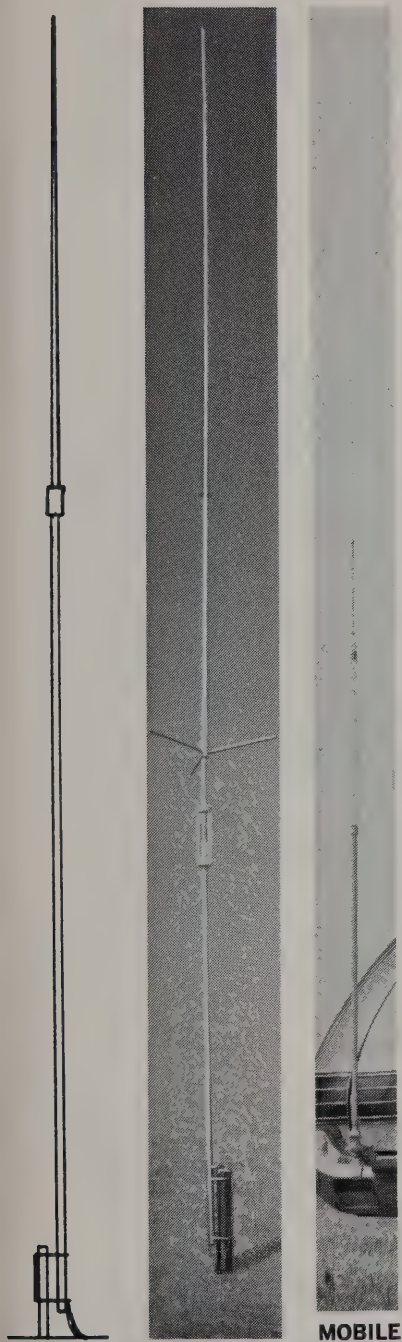
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high-performance balanced mixer for 2304 MHz

Construction details for a hot-carrier diode mixer using stripline design that can provide a 7-dB noise figure at 2304 MHz

Paul C. Wade, WA2ZZF, 153 Woods Road, Somerville, New Jersey 08876

In a previous article about a 1296 MHz balanced mixer[1] I mentioned using a similar circuit for 2304 MHz. This elicited several requests for information about the 2304 MHz mixer shown in **fig. 1**. However, since I felt that the ceramic board construction was beyond the facilities available to most amateurs, I was unable to respond. Since I was curious to see if simple construction techniques would still work at 2304 MHz, I finally made the version shown in **fig. 2** which uses an ordinary printed-circuit board. To my surprise, the second version exhibits *better* performance than the "professional" one.

The basic design, shown schematically in **fig. 3**, is identical to the 1296 MHz mixer described in *QST* and consists of a 3 dB quadrature-hybrid coupler, quarter-wave stubs for bypassing, and a low-Q pi network for i-f impedance matching. The major difference is the use of 1/32 inch (0.8mm) double-clad, G10 epoxy-fiberglass circuit board, rather than 1/16 inch (1.5mm), to maintain a reasonable aspect ratio. The line dimensions shown in **fig. 4**, however, are more critical because of the thinner board and higher frequency.

I cut the printed-circuit mask directly to size on Rubylith* with a knife and a ruler so it should be possible to

duplicate the layout with tape or by cutting the pattern directly into the copper and peeling away the excess. Dimensions A through E, and especially line widths A and B, should be within 0.005 inch (0.1mm) of the values shown for best results.

construction

Construction is exceedingly simple, requiring only a drill and a vise. The circuit board is attached to a shelf in a 2-3/4 x 2-1/8 x 1-5/8 inch (70x54x41mm) Minibox. The shelf dimensions are shown in **fig. 5**. Use enough screws to keep the board flat against the shelf and to provide a ground path for capacitor C4. Metal screws have no effect if they are kept away from the microstrip transmission lines. Approximate placement of the screws can be seen in **fig. 2**.

Critical parts are the blocking capacitors, C1 and C2, the connectors, and, of course, the mixer diodes. The blocking capacitors should be low-loss chip capacitors, preferably physically small to maintain a low vswr. The same considerations, loss and vswr, also apply to the connectors; SMA type connectors similar to the ones shown are available very reasonably from E. F. Johnson.

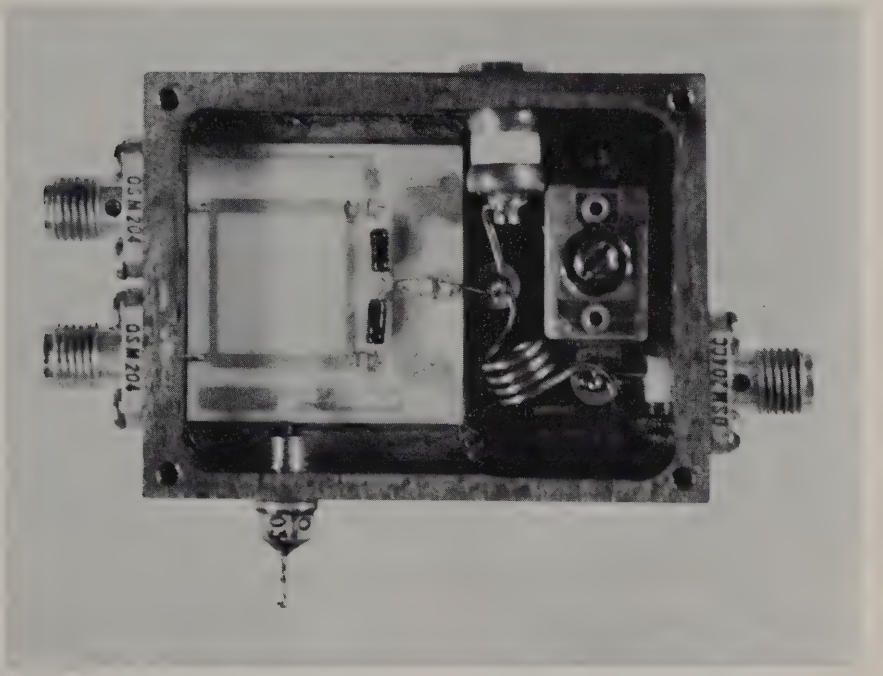


fig. 1. Original balanced mixer for 2304 MHz used ceramic substrate. Small blobs are silver paint used for fine tuning.

adjustment

Adjustment is the height of simplicity. A two-meter converter is connected to the i-f output, and a *clean* 1 to 2 milliwatts at 2160 MHz is fed into the LO connector. Apply about 1.5 mA of bias current to the diodes and apply a

*"Rubylith" is a trademark of Ulano Co.

moderately strong signal at 2304 MHz. When the bias current and i-f trimmer capacitor are adjusted for maximum signal, the tune-up is completed.

performance

A testimonial to the performance of this mixer is that, at the Eastern VHF/UHF Conference earlier this year, the original version won a certificate for the lowest noise-figure 2304 MHz converter. The measured noise figure was 7.1 dB (i-f noise figure, 2 dB), but, in the interest of fairness, W1JAA insisted that 3 dB be added because of the poor image rejection, so the final noise figure was 10.1 dB. The new version described in this article measures *2 dB better*! The mixer has a conversion loss of 6.1 dB.

Noise figure, both theoretical and measured, is the sum of conversion loss plus i-f noise figure. With a 1 dB noise figure at two meters, an overall noise figure of approximately 7 dB could be realized. This should compare favorably with most other mixers at this frequency; W2CQH's interdigital mixer[2] may be better, but it is also more complex. Isolation between the local-oscillator to the rf ports measures 22 dB; this indicates that the hybrid coupler is working properly and that the diodes are well matched to 50 ohms (any power reflected from the diodes will decrease the isolation).

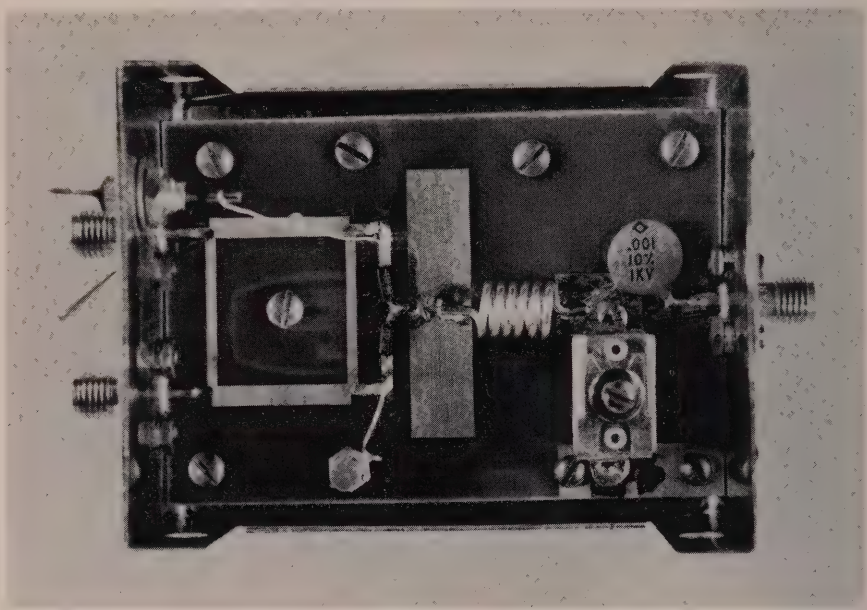
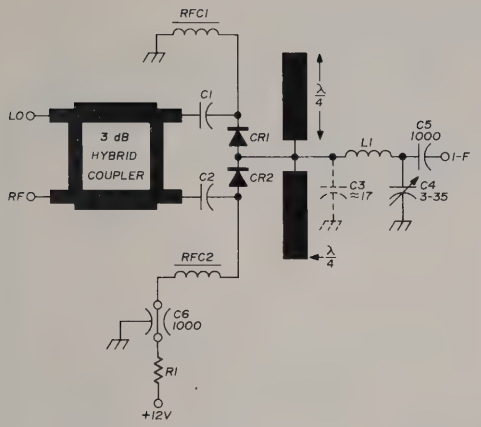


fig. 2. Balanced mixer for 2304 MHz using G10 printed-circuit board. No tuning is necessary.

dc bias

A few words about the value of dc bias in mixers may be in order. I have not seen any other amateur designs which include biasing of the mixer diodes, whereas it is fairly common practice in commercial mixers. Prior to building my first balanced mixer (for 1296 MHz), I made some measurements of diode impedance versus rf signal power (LO) and bias current. The conclusion was that, with dc bias, less



C1,C2	50 to 1000 pF chip capacitor (not critical, see text)
C3	stray capacitance of quarter-wavelength stubs (≈17 pF)
C4	3-35 pF mica trimmer (Arco 403)
C5	1000 pF disc ceramic
C6	1000 pF feedthrough bypass
CR1, CR2	hot-carrier diodes rated for 3 GHz operation (H-P 5082-2535, H-P 5082-2565, Alpha D5501, matched pair not essential)
L1	6 turns no. 18 (1.0mm), 3/8" (9.5mm) long, wound on 3/16" (5mm) mandrel
R1	5k to 15k, vary to set diode bias current
RFC1,2	1 turn no. 30 (0.25mm), wound on 1/16" (1.5mm) mandrel
J1-J3	SMA connector (OSM 215 or E. F. Johnson 142-0297-001)

fig. 3. Schematic diagram of the balanced mixer for 2304 MHz. Dc bias is provided to the diodes to minimize LO requirement. Full-size printed-circuit layout is shown in fig. 4.

LO power is required to raise the diode impedance to 50 ohms, and the impedance is less sensitive to the drive level.

We are, in essence, substituting readily available dc power for hard-to-get rf power! This is borne out by the fact that my mixers work fine at LO levels of one milliwatt, while other, similar designs[3,4] specify around three milliwatts. Also, Tilton recently mentioned[5] several balanced mixers which were rejected by *QST* — their poor performance was apparently due to lack of LO power. Addition of dc bias might have helped significantly.

Bias current is not critical (in this mixer minimum conversion loss was achieved with 1.8 mA of diode current), but it can be varied from 1.0 to 2.6 mA with only a 1 dB increase in loss. With no bias, however, conversion loss increased to 13 dB. Small changes in LO power, simulating normal drift, also had a minimal effect.

One final advantage of dc biasing may be to force the diodes to operate at the same current, and hence at similar impedance, for no attempt at diode matching or individual tuning was made in order to achieve the stated performance.

balanced-mixer design

I have received several queries about designs for other frequencies. These can be made by direct scaling from this design or the 1296 MHz version, following these guidelines:

1. Dimension A is for a characteristic impedance, Z_o = 50 ohms, and B for Z_o = 35 ohms. No change is required if the same board material is used.

2. Dimensions C and D are one-quarter wavelength long at a frequency halfway between the signal and LO frequencies (2232 MHz in this case).

3. Dimension F is one-quarter wavelength long at the signal frequency (2304 MHz).

4. Dimension G is one-quarter wavelength long at the local-oscillator frequency (2160 MHz).

5. Wavelength in microstrip line is a function of Z_o, so use a graph of Z_o and λm if you change board material (different dielectric constant).

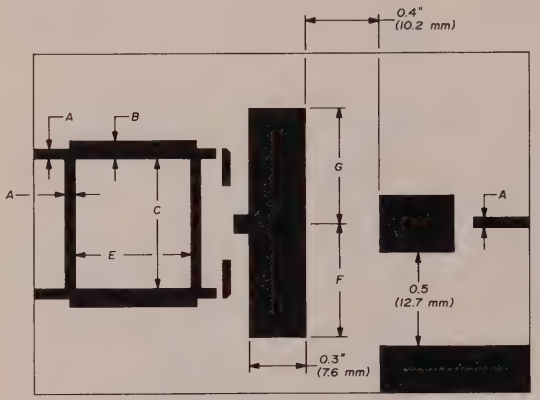
6. Dimension E is D minus A.

7. Use a low Q i-f circuit which matches the impedance of the diodes (100 to 200 ohms at the i-f).

If the same board material is used (1/32 inch G10) scaling dimensions C, D, F and G directly by the ratio of the frequencies will work fine. As mentioned above, good isolation from the LO to rf ports is an indication that everything is working properly; measure this after all other adjustments have been made.

conclusion

The balanced mixer will provide excellent performance which is easily duplicated and maintained. Use of a low-loss filter at the input is recommended to improve image rejection; there are no other responses if a clean local-oscillator signal is used. A preamplifier may be used for even lower noise figure, and the mixer will provide a low vswr load for the preamp to help stability. However, unless you really need the small (and costly) improvement a preamp provides, the mixer alone has two major advantages — it is much harder to burn out, and it will *not* oscillate.



dimensions		
A	0.053"	(1.3mm)
B	0.096"	(2.4mm)
C	0.696"	(17.7mm)
D	0.678"	(17.2mm)
E	0.625"	(15.9mm)
F	0.610"	(15.5mm)
G	0.650"	(16.5mm)

fig. 4. Full-size printed-circuit board for the 2304-MHz balanced mixer. Microstrip dimensions shown at right may be scaled to other frequencies (see text).

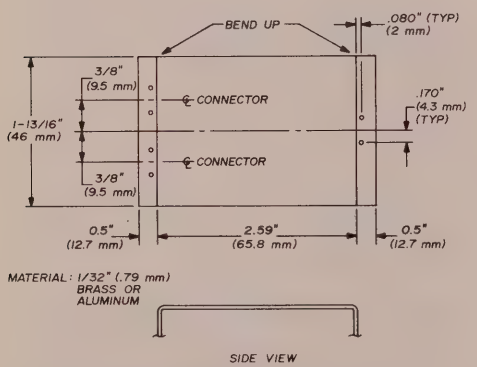


fig. 5. Shelf dimensions for installing the balanced mixer in a standard Bud CU-3000A Minibox.

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satellite receivers for repeaters

Adding satellite receivers to your repeater won't bring in OSCAR 7 — but it might improve your terrestrial coverage

Fred Studenberg, Jr., WA4YAK*

Many users of low-powered handie-talkies have found that the theory of reciprocity proves to limit their ability to communicate through a repeater. Reciprocity implies that, with equal receiver sensitivity and antenna gain on both ends of a path, equal transmitter power is necessary for both stations to hear each other with the same signal strength.

Generally, a repeater's transmitter will have a 17 to 23 dB power advantage over a two-watt portable, effectively reducing the repeater's usable sensitivity by the same amount. This is the typical case of hearing the repeater full quieting but not being able to access it, even though the repeater's receiver is very sensitive.

Assuming that it is desired to retain the present transmitter coverage, there are several ways to balance out the transmit-receive range. In the case of a split-site system, the existing receiver can be relocated at the highest elevation possible. This is, of course, assuming that the repeater was set up initially with equal transmitter and receiver

*Electronics Communications, Inc., St. Petersburg, Florida 33710

heights. Many repeaters were originally designed for use with mobiles that ran 30 to 60 watts and tended to hear as well as they transmitted to the repeater.

satellite receivers

If you operate a single-site system, or have run out of tall buildings and towers, there is another solution to the problem. The concept of satellite receivers has been applied successfully for some years by commercial users and some advanced repeater groups. These remote receivers, located in areas away from the main receiver, relay the weaker incoming signals by vhf or uhf links to the main transmitter.

Ideally, a satellite receiver system should have circuitry that will automatically select the audio from the receiver that is hearing the best quieting signal. However, our repeater group has found that very elaborate methods are not really needed. A typical fm receiver will yield usable audio from any signal that can open the squelch, even though the receiver may only be quieted by 10 or 15 dB. With this in mind, a four-channel scanner receiver was built for the 450-MHz band. **Fig. 1** shows how this is used in our satellite system.

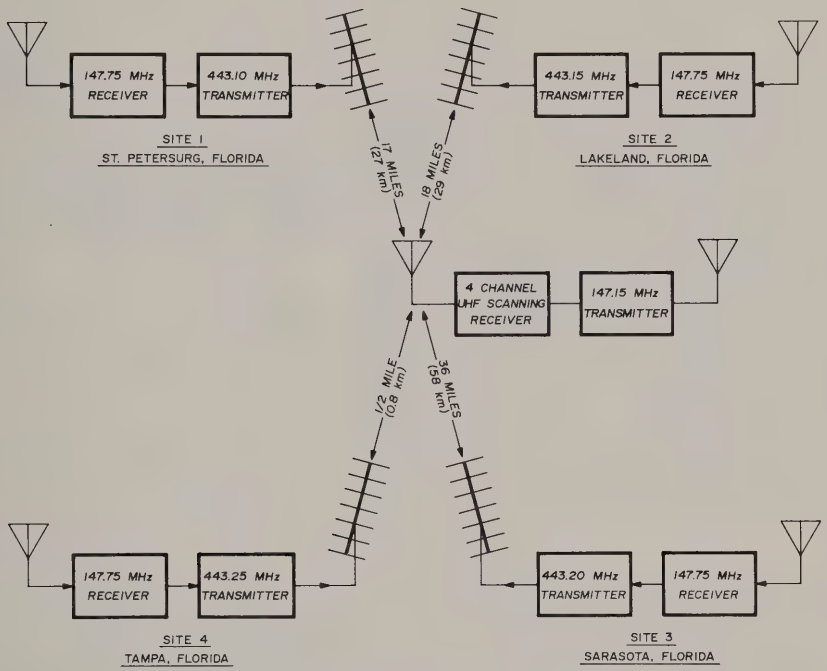


fig. 1. Satellite receivers expand the two-way coverage area of repeater WR4ALT in Tampa, Florida.

The receiver will lock up on the first link transmitter it hears during the scan process. The system tends to seek out the link transmitter that is carrying the best signal, since a link carrying a "chopping" mobile signal will also be "chopping" into the link receiver. The scanner link receiver will tend to unlock and lock up on a solidly keyed link

from a receiver that is hearing a full quieting signal. To avoid wear on the link transmitter's keying relay and also to provide security to the link receiver, the actual presence of a signal is indicated by a 10-kHz subcarrier modulated on the link transmitter. The squelch, and consequently the scanning action, is dependent upon the presence of the 10-kHz subcarrier.

problems encountered when trying to make one-to-one comparisions in strengths of the remote receivers.

To satisfy FCC requirements, each satellite system is licensed as an auxiliary link station, and the link transmitters are identified continuously by a 4-kHz MCW signal. This is filtered out in the link receiver, and the main transmitter is identified by its normal identi-

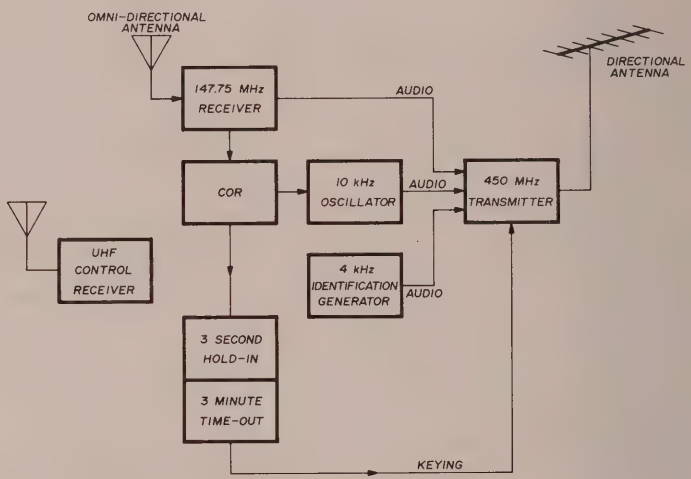


fig. 2. A typical remote receiver installation. The 10-kHz sub-carrier ensures that 450-MHz signals from other sources besides the satellite receivers will not be repeated on the 147-MHz output frequency. It also serves as a "tailless" squelch since the 450-MHz transmitter is held on by the 3-second hold timer.

Referring to **fig. 2**, the satellite receiver's COR keys the 10-kHz oscillator and also the link transmitter, but through a three-second hold-in timer. This effectively eliminates the squelch tail from the link receiver since it resumes scanning when the subcarrier is removed, even though the carrier is still present. A 10-kHz tone was chosen to allow fast detection time. A Signetics NE567 phase-locked loop tone decoder is used, and it can recognize the subcarrier in less than 1 millisecond. A scan search rate of 25 milliseconds per channel is used, allowing a worse case delay of about 80 milliseconds when selecting different satellite receivers. This has proven to be a very acceptable method of voting without any of the fication circuits. Regulations require control of the link transmitter. In our case, a common radio control frequency is used for all five sites, simplifying the control operator's equipment requirements. Link frequencies in the 450-MHz band were chosen since monitoring of the link transmitters is not required. Link operation on 220 MHz would require four-frequency monitoring capability.

Selection of the remote sites should be dictated by "dead spots" in the repeater's current coverage area. Naturally, high antenna heights are desirable, but even low antenna satellite receivers will do wonders for filling in receive coverage for the handie-talkie users.

ham radio

crystal discriminator for vhf fm

High-performance crystal discriminator for vhf fm can be built around an inexpensive third-overtone CB crystal

G. Kent Shubert, WAØJYK, 1308 Leeview Drive, Olathe, Kansas 66061

Many amateurs, and professional design engineers as well, have attempted to design single-conversion fm receivers only to scrap their brainchild because of low recovered audio and the resulting noisy audio output with poor squelch action. A single-conversion vhf fm receiver must have an intermediate frequency of 5 MHz or higher to provide adequate image rejection. The percentage of deviation is quite small at the higher intermediate frequencies so ICs utilizing quadrature detection may provide less than a millivolt of recovered audio. A simple solution to this problem is to increase the Q of the quadrature coil or use a crystal in its place. Unfortunately, however, these ICs become very unstable and impossible to tame when the Q of the quadrature resonator is increased.

crystal discriminators

Most of the *novel* discriminator circuits, such as the pulse-counting or digital type and transformerless type, as well as the conventional Foster-Sealey, Round-Travis and the ratio detector, do not provide enough recovered audio to be useful at the higher i-f frequencies. Crystal discriminators are quite popular in some of the more sophisticated commercial radios. In fact, Motorola uses a two-crystal discriminator to obtain a plus and minus voltage swing for automatic frequency control in some radios. The crystal discriminator was mentioned briefly in *QST* but no specific values were given.[1]

Most crystal filter manufacturers market a crystal discriminator which

complements their crystal filter line so they are very reluctant to disclose the internal workings of these devices. However, several hours of investigation turned up a specification sheet from C-F Networks, which included all the necessary data to design other crystal discriminators and tune them.

Excellent 9-MHz crystal filters (manufactured by KVG) are available from Spectrum International.* In addition, monolithic crystal filters and monolithic filter elements are manufactured by Piezo Technology, Inc.† Spectrum International also markets an excellent

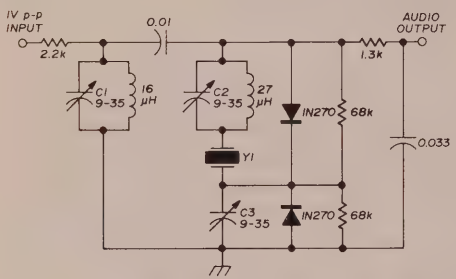


fig. 1. Low cost fm crystal discriminator uses third-overtone channel-1 CB crystal (26.965 MHz) at its 9-MHz fundamental. Response curve of this circuit is plotted in fig. 2.

KVG crystal discriminator for approximately $25.

Quite a few CB crystals are available on the surplus market and in various ham-shack junkboxes so one of these 27 MHz, third-overtone crystals was put into a breadboard circuit and carefully tested at its fundamental frequency, 9 MHz, for modulation acceptance, recovered audio and distortion. Modulation acceptance was sufficient for 5 kHz deviation but more than 7 kHz deviation caused slight audible distortion. A

*Spectrum International, Box 1084, Concord, Massachusetts 01742.

†Piezo Technology Inc., Post Office Box 7877, Orlando, Florida 32804.

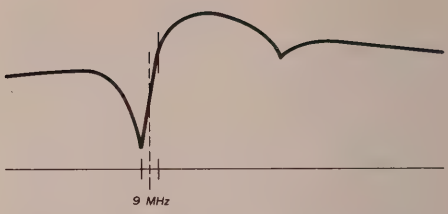


fig. 2. Response of the 9-MHz crystal discriminator. The linear section of the response curve just above the fundamental resonant frequency is the portion actually used. Good filtering is required ahead of this discriminator or serious audio distortion will result.

12 kHz bandpass filter would provide a perfect match for the discriminator.

The crystal used for the tests was a channel-1 transmit crystal for a Heathkit Model CB-1. It is a 26.965 MHz, third-overtone crystal identical to those used in hundreds of other CB radios. This particular crystal is in the larger HC-6/U holder but many are in smaller holders.

the circuit

Fig. 1 is the schematic of the crystal discriminator circuit. Capacitor C3 is adjusted for zero voltage with an unmodulated carrier at center frequency but this setting may not hold for all crystals, and it is possible to obtain good performance and good audio re-

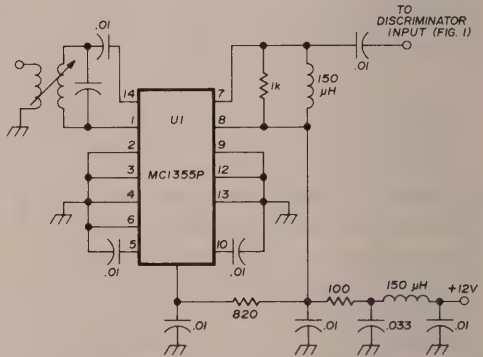


fig. 3. Amplifier/limiter for driving the crystal discriminator uses a Motorola MC1355P intergrated circuit.

covery from crystals that will not tune for a zero center. This just means that the crystal doesn't provide a good linear portion at exactly 9 MHz but with inexpensive, surplus crystals you can't be too particular.

Capacitors C1 and C2 are most easily adjusted with an audio-frequency sine wave applied to an fm signal generator or transmitter and using an oscilloscope to check distortion of the recovered audio sine wave. This same method should be used to align quadrature detectors, too, since there is no true zero-center reading. With a 1 volt p-p i-f signal (at 9 MHz) and 5 kHz deviation, the recovered audio will be about 1 volt p-p at the lower audio frequencies, rolled off at the higher frequencies with the de-emphasis network. The de-emphasis circuit also removes most of the remaining 9 MHz from the audio.

This type of crystal discriminator is amplitude sensitive and requires a good limiter ahead of it to provide a-m rejection. The slope of the discriminator is shown in **fig. 2** but may vary a little with different crystals. It is possible to detect fm on the descending portion of the curve about 40 kHz lower but modulation acceptance is only a few kHz.

The Motorola MC1355P is an excellent IC to drive this discriminator. **Fig. 3** shows a possible circuit, but the MC1355P has more than 60 dB gain available so it requires very careful layout, bypassing and shielding.

If you're not a reformed CBer it's possible to cultivate a friendship with a CBer who has recently acquired a vfo or you may opt to purchase a new crystal from JAN Crystals* for $2.50.

*JAN Crystals, 2400 Crystal Drive, Fort Myers, Florida 33901.

reference

1. Douglas A. Blakeslee, W1KLK, "Receiving FM, Part II," *QST*, February, 1971, page 16.

ham radio

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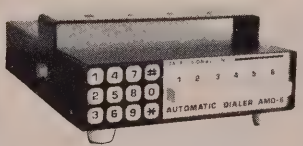
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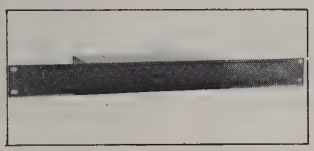
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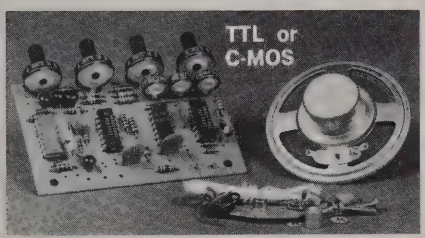
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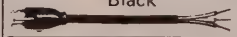
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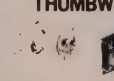
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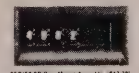


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SERIES SF Front Mount Assembly (S11 78)

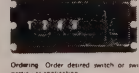


Note: Picture to the left has the following
4 ea. SF 12, 1 ea. SF EP, 1 ea. SF BB
1 ea SF DP and 1 ea SF HB

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SF EP	End Plate (Pair)	.50
SF DP	Divider Plate (each)	.40
SF BB	Blank Body (each)	.40
SF HB	Half Body (each)	.40

SERIES SR Rear Mount Assembly (S12 78)



Note: Picture to the left has the following
4 ea. SR 12, 1 ea. SR EP, 1 ea. SR BB
1 ea SR DP and 1 ea SR HB

ACCESSORIES

Part No.	Description	Price
SR EP	End Plates (pair)	$ .50
SR DP	Divider Plate (each)	.40
SR BB	Blank Body (each)	.40
SR HB	Half Body (each)	.40

Ordering: Order desired switch or switches and add necessary accessories for your particular application.

NEW PROTO BOARD-100



Here's a low cost, big 10 IC capacity breadboard kit with all the quality of QT Sockets and the best of the Proto Board series . . . complete down to the last nut, bolt and screw. Includes 2 QT-35S Sockets; 1 QT-35B Bus Strip, 2 5-way binding posts; 4 rubber feet; screws, nuts, bolts; and easy assembly instructions

COMPLETE KIT... *Special $17.95

DISPLAY LEDS

FND70	Com. Cath.	.250	.69
MAN 1	Com. Ano.	.270	$1.95
MAN 2	5x7 Matrix	.300	3.95
MAN 3	Com. Cath.	.125	.39
MAN 4	Com. Cath.	.187	1.95
MAN 7	Com. Ano.	.30	1.50
DL33	Com. Cath.	.125	1.95
DL747	Com. Ano.	.625	2.50

DISCRETE LEDS



MV 10	Red	5/$1.00
MV 50	Red	6/$1.00
MV 5024	Red	5/$1.00
MV 5024	Green	4/$1.00
MV 5024	Yellow	4/$1.00
MV 5024	Orange	4/$1.00
MV 50*	Red 1¼" Leads	5/$1.00

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5 function plus constant – addressable memory with individual recall – 8 digit display plus overflow – battery saver – uses standard or rechargeable batteries – all necessary parts in ready to assemble form – instructions included. 3" x 5¼" **SPECIAL $12.95 each**

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MODEL 100Ω 500Ω 1K
2K 5K 10K 20K
100K 200K 1Meg

SPECIAL *Model TR11*

RESISTANCES (OHMS)	1–9	10–24
STD 100Ω – 1 MEGΩ	4/$1.00	

1/16 VECTOR BOARD



P Pattern
0.1" Hole Spacing

MATERIAL	STOCK NO	LONG	WIDE	PRICES 1	PRICES 20
PHENOLIC	84P44 062XXXP	4.50	6.50	1.72	1.54
	169P44 062XXXP	4.50	17.00	3.86	3.32
EPOXY GLASS	84P44 062	4.50	6.50	3.01	1.86
	84P44 062 -	4.50	8.50	2.96	2.31
	169P44 062	4.50	17.00	5.04	4.53
	169P54 062	8.50	17.00	9.21	8.26
EPOXY GLASS COPPER CLAD	169P44 062C1	4.50	17.00	6.80	6.12

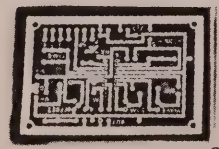
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12 or 24 Hour — 115 Vac
25' VIEWING DISTANCE
Walnut Case-6" x 3" x 1"
Hr. & Min.-6" High
Seconds-3" High
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IC'S EXAR KITS

FUNCTION GENERATOR KIT



features sine, triangle and square wave, THD 0.5% typ., AM/FM capability

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Includes monolithic function generator IC, PC board, and assembly instruction manual.

XR-2206KB — **SPECIAL $27.95**
Same as XR 2206KA above and includes external components for PC board.

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XR-2240CP	Programmable Counter/Timer	**SPECIAL 3.25**
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14 pin	.20	.19	.18	28 pin	.45	.44	.43
16 pin	.22	.23	.22	36 pin	.60	.59	.58
18 pin	.29	.28	.27	40 pin	.63	.62	.61
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16 pin	.33	.31	.29	36 pin	1.39	1.26	1.15
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18 pin	.75	.68	.62	40 pin	1.75	1.55	1.40

50 PCS. RESISTOR ASSORTMENTS $1.75 PER ASST.

ASST. 1	5 ea:	10 OHM- 27 OHM	12 OHM- 33 OHM	15 OHM- 39 OHM	18 OHM- 47 OHM	22 OHM- 56 OHM			1/4 WATT 5% = 50 PCS.
ASST. 2	5 ea:	68 OHM- 180 OHM	82 OHM- 220 OHM	100 OHM- 270 OHM	120 OHM- 330 OHM	150 OHM- 390 OHM			1/4 WATT 5% = 50 PCS.
ASST. 3	5 ea:	470 OHM- 1.2K	560 OHM- 1.5K	680 OHM- 1.8K	820 OHM- 2.2K	1K- 2.7K			1/4 WATT 5% = 50 PCS.
ASST. 4	5 ea:	3.3K 8.2K	3.9K 10K	4.7K 12K	5.6K 15K	6.8K 18K			1/4 WATT 5% = 50 PCS.
ASST. 5	5 ea:	22K 56K	27K 68K	33K 82K	39K 100K	47K 120K			1/4 WATT 5% = 50 PCS.
ASST. 6	5 ea:	150K 390K	180K 470K	220K 560K	270K 680K	330K 820K			1/4 WATT 5% = 50 PCS.
ASST. 7	5 ea:	1M 2.7M	1.2M 3.3M	1.5M 3.9M	1.8M 4.7M	2.2M 5.6M			1/4 WATT 5% = 50 PCS.

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IN751A	5.1	400m	4/1.00	IN4004	400 PIV	1 AMP	.10
IN752	5.6	400m	4/1.00	IN3600	50	200m	6/1.00
IN753	6.2	400m	4/1.00	IN4148	75	10m	15/1.00
IN754	6.8	400m	4/1.00	IN4154	35	10m	12/1.00
IN965B	15	400m	4/1.00	IN4734	5.6	1w	.28
IN5232	5.6	500m	.28	IN4735	6.2	1w	.28
IN5234	6.2	500m	.28	IN4736	6.8	1w	.28
IN5235	6.8	500m	.28	IN4738	8.2	1w	.28
IN5236	7.5	500m	.28	IN4742	12	1w	.28
IN456	25	40m	6/1.00	IN4744	15	1w	.28
IN458	150	7m	6/1.00	IN1183	50 PIV	35 AMP	1.60
IN485A	180	10m	5/1.00	IN1184	100 PIV	35 AMP	1.70
IN4001	50 PIV	1 AMP	.09	IN1186	200 PIV	35 AMP	1.80
IN4002	100 PIV	1 AMP	.10	IN1188	400 PIV	35 AMP	3.00

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MPS-A06	5/$1.00	PN3567	3/$1.00	PN4250	4/$1.00
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2N2369	5/$1.00	2N3705	5/$1.00	2N4403	4/$1.00
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10pf	.05	.04	.03	.001	.05	.04	.035
22 pf	.05	.04	.03	.0047	.05	.04	.035
47 pf	.05	.04	.03	.01	.05	.04	.035
100 pf	.05	.04	.03	.022	.06	.05	.04
220 pf	.05	.04	.03	.047	.06	.05	.04
470 pf	.05	.04	.035	.1	.12	.09	.075

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.001mf	.12	.10	.07	.022mf	.13	.11	.08
.0022	.12	.10	.07	.047mf	.21	.17	.13
.0047mf	.12	.10	.07	.1mf	.27	.23	.17
.01mf	.12	.10	.07	.22mf	.33	.27	.22

+ 20% DIPPED TANTALUMS (SOLID) CAPACITORS

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.15 35V	.28	.23	.17	2.2 25V	.31	.27	.22
.22 35V	.28	.23	.17	3.3 25V	.31	.27	.22
.33 35V	.28	.23	.17	4.7 25V	.32	.28	.23
.47 35V	.28	.23	.17	6.8 25V	.36	.31	.25
.68 35V	.28	.23	.17	10 25V	.40	.35	.29
1.0 35V	.28	.23	.17	15 25V	.63	.50	.40

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	Axial Lead				Radial Lead			
.47	50	.15	.13	.10	47 25	.15	.13	.10
1	50	.16	.14	.11	47 50	.16	.14	.11
3.3	50	.15	.13	.10	1 16	.15	.13	.10
4.7	25	.16	.14	.12	1 25	.16	.14	.11
10	25	.15	.13	.10	1 50	.16	.14	.11
10	50	.16	.14	.12	4.7 16	.15	.13	.10
22	25	.17	.15	.12	4.7 25	.15	.13	.10
22	50	.24	.20	.18	4.7 50	.16	.14	.11
47	25	.19	.17	.15	10 16	.14	.12	.09
47	50	.25	.21	.19	10 25	.15	.13	.10
100	25	.24	.20	.18	10 50	.16	.14	.12
100	50	.35	.30	.28	47 50	.24	.21	.19
220	25	.32	.28	.25	100 16	.19	.15	.13
220	50	.45	.41	.38	100 25	.24	.20	.18
470	25	.33	.29	.27	100 50	.35	.30	.28
1000	16	.55	.50	.45	220 16	.23	.17	.15
2200	16	.70	.62	.55	470 25	.31	.28	.25

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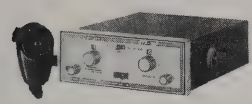
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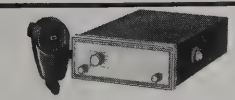
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2-meter FM, 100 channels, 30 watts (incl. 146.94 MHz)
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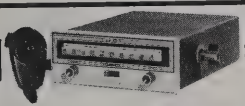
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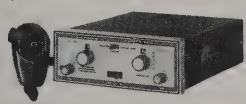
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GTX-2
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NAME_____ AMATEUR CALL_____

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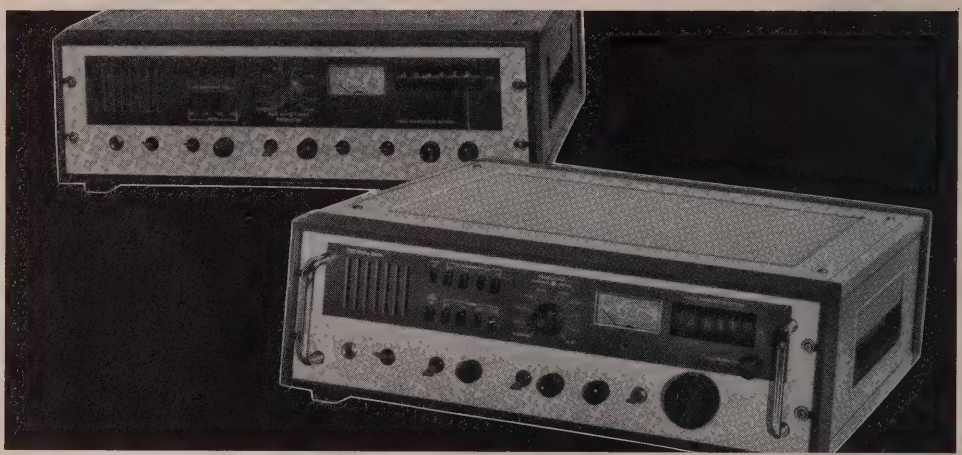
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Raleigh, North Carolina 27611

Please send complete FREE information on the exciting new:

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CITY _____ STATE _____ ZIP _____

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Federal Supply Schedule Group 58 Part VII, Contract GS-OOS-24016

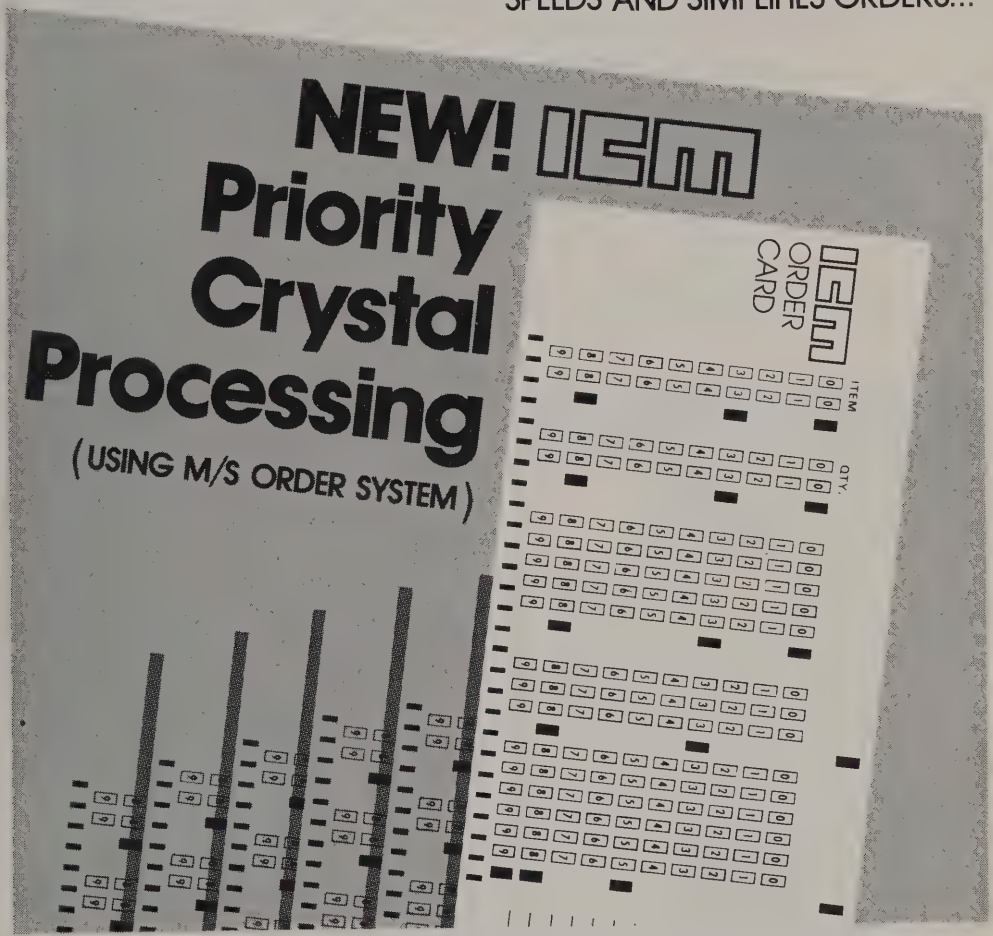
ITT Mackay Marine

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Our new priority crystal processing, using a mark sensing order system, is designed to expedite orders for International Crystals and EX Kits. It's another effort to improve our order processing and make a proven reliable system even better. Your order in the future will be processed by using mark sensing order cards and prepunched name and address cards. Complete and return the coupon and we will mail you a special kit containing instructions, order cards and a prepunched name and address card. This is the first step in our new "no delay" order processing.

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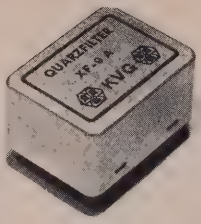
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IF Freq. †	28-32	28-32	28-32	28-32	28-32
N.F. (typical)	2.5dB	2.8dB	3.4dB	3.8dB	9.0dB
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With Reed Relay, add $3.00; Wired & Tested, add $8.45
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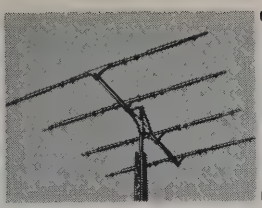
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HOT COMBO



Element for element, our rugged High-Q beam antennas are designed to give you the biggest signal your transmitter is capable of.

Now when you install one of these Swan antennas you can make sure you're running full bore all the time by hooking our new SWR/RF Power meter in the coax.

Heavy duty 4-element Tribander Four elements work on 10, 15 and 20 meters. Optimum spacing for maximum performance. Precision tuned, weatherproof traps. 100-mph winds. TB-4HA. *$249.95*

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SWR/RF Power meter Combination meter measures standing wave ratio **and** antenna power. Low insertion loss lets you leave it in circuit. 3.5 to 150 MHz. *$21.95*

All Swan Beam Antennas are Rated for 2000 Watts and designed to use 52 Ohm coaxial feedlines.

SWAN BEAM ANTENNA SPECIFICATIONS

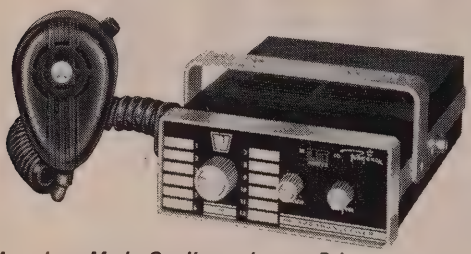
Antenna Model Number	Average Forward Gain	Front to Back Ratio	Boom Length & Diameter	Longest Element	Turning Radius	Maximum Wind Survival	Wind Load @ 80 mph	Wind Surface Area	Net Weight
TB-4HA	9 dB	24-26 dB	24' x 1.5"	28'-10"	18'-6"	100 mph	148 lbs.	6 sq. ft.	54 lbs.
TB-3HA	8 dB	20-22 dB	16' x 1.5"	28'-2"	16'	100 mph	110 lbs.	4 sq. ft.	44 lbs.
TB-2A	5 dB	16-18 dB	6.5' x 1.5"	27'-8"	14'-3"	80 mph	60 lbs.	1.8 sq. ft.	18 lbs.
MB-40H	4 dB	16-18 dB	15.75' x 1.5"	30'-4"	17'-6"	100 mph	80 lbs.	2.5 sq. ft.	40 lbs.

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The HAL DKB-2010 Dual Mode keyboard is one of the most sophisticated products ever offered to the radio amateur. It's an all solid state keyboard that allows you to send either RTTY or CW — with more ease, more versatility than anything you've ever seen before.

In the RTTY mode, you can transmit at standard data rates of 60, 66, 75 or 100 WPM, as well as an optional 132 WPM, 100 baud. In addition to the complete alphanumeric keys, you get 17 punctuation marks, 3 carriage control keys, 2 shift keys, a break key, 2 three-character function keys, a "DE-call letters" key and a "Quick brown fox . . ." test key.

In the CW mode, you can send at speeds anywhere between 8 WPM and 60 WPM. You can also adjust dot-to-space weight ratios to your liking. For CW, you have all alphanumeric keys, plus 11 punctuation marks, 5 standard double-character keys, 2 shift keys, a break-for-tuning key, error key, "DE-call letters" key, plus 2 three-character function keys. Output interfacing is compatible with cathode keying or grid-block keying. A side tone oscillator and built-in speaker allow you to monitor your signal — with adjustable volume and pitch controls.

The DKB-2010 also has a three-character memory buffer which operates in either the RTTY or CW mode, allowing you to burst type ahead without losing characters. A 64-character memory buffer is also available as an option. Key function logic in either mode is governed by LSI/MOS circuitry. All key switches are computer grade.

The DKB-2010 is available assembled or in kit form. Should you choose the kit, you'll find construction easy — the unit consists of three assemblies: power supply board, logic PC board, keyswitch PC board, and pre-assembled wiring harness.

Any way you look at it — as an easy-to-build kit, a complete assembly, as a CW keyboard, or an RTTY keyboard, the HAL DKB-2010 is a real breakthrough for every amateur. It adds a whole new dimension to the exciting world of amateur radio. Once you've used the DKB-2010, you'll wonder how you ever got along without it!

Prices: $425 Assembled; $325 Kit

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Telephone: (217) 367-7373

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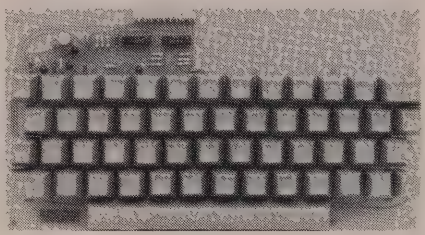
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* 53 Keys
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We are happy to announce a new addition to our keyboard and encoder line. Our new KBD-3 uses a one chip MOS encoder system to give you maximum possible features with a minimum number of parts.

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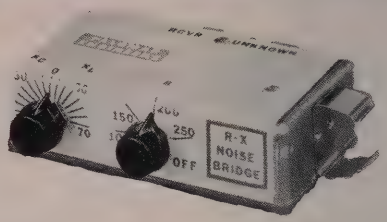
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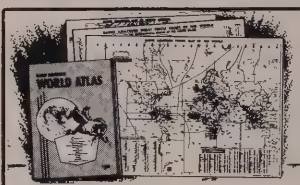
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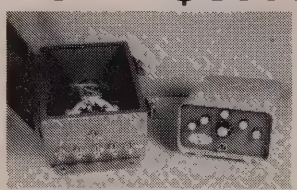
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The SW-5 is a remote controlled RF switch with indicator lights telling which antenna is in use. It will handle 4 kW PEP and more. Remote switch is housed in weather tight hinged box. A six wire #18 cable is required to operate the SW-5. Ham M control cable works fine up to 150′. Heavier cable necessary for longer distances. Remote switch operates off 28 VDC built in power supply. No visible effects on SWR. Zero dB insertion loss. Not recommended above 30 MHz. Standard unit is equipped with UHF connectors but BNC, N, HN, C connectors are available at additional charge. Models available are SW3, 4, 5, 6, 7, 8, 9. Also heavy duty 10kW units. Special switching systems are available. Tell us your needs.

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The Transceiver you'd expect in 1980



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The benefits of ALC now extend to output powers less than full rating with a front panel threshold control. When driving linears that require less than maximum available power from the TRITON, or when propagation conditions permit reliable contacts at reduced power levels, ALC will hold your output to the desired level.

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And to top it all off, the highly desirable case geometry has been maintained, but it has a handsome new look. Bold lettering on an etched aluminum front panel and textured black sides and top make the TRITON IV look as sharp as it performs.

There is nothing like a TRITON IV for reliability, features, value and just pure fun. And — best of all — you do not have to wait until 1980 to own one.

TRITON IV $699.00

For more information about the new TRITON, as well as the full line of accessories that will be available soon, see your dealer or write.

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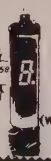
3DB Bandwidth: 220 KHZ
50DB Bandwidth: 700 KHZ (Max)
Insertion Loss : 10 DB (Max)
Impedance : 330 ohms

The bandwidth can be narrowed by loading filter with a higher impedance

RCA CA3043
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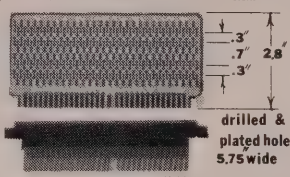
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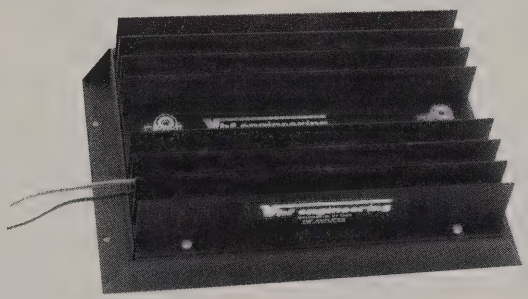
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- All solid state
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PA140/10 5-15 watts in for 100-150 watts out @ 13.6 volts
Typically 140 watts for 10 watts in

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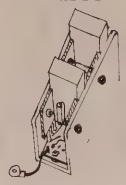
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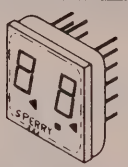
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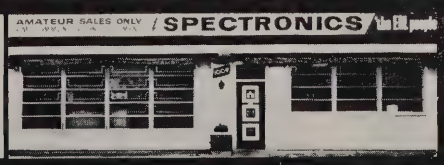
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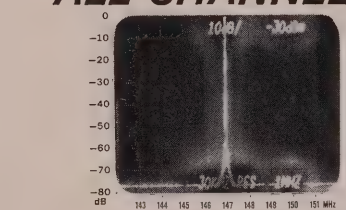
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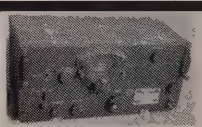
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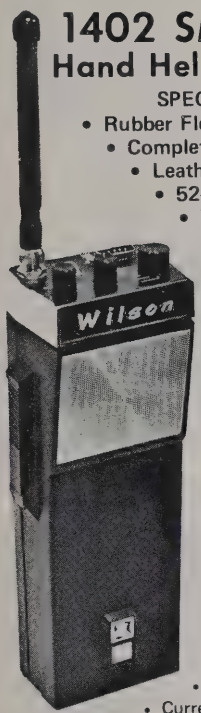
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Hand Held FM Transceiver



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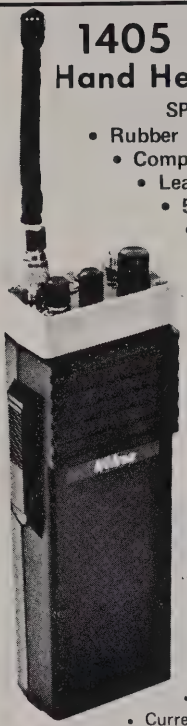
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- 10.7 IF and 455 KC IF.
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- Current Drain RX 14 MA TX 450 MA.
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TE-1	Continuous Sub-Audible Tone Encoder (Installed)	$ 39.95
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- Microswitch Mike Button.
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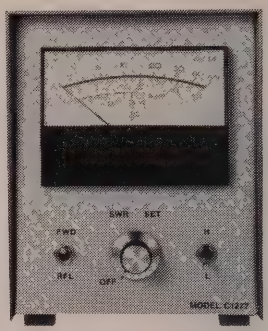
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CW	50 Watts 100-225 MHz
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R-4C FEATURES:

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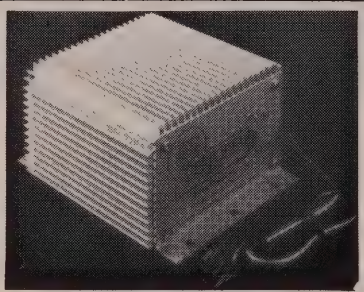
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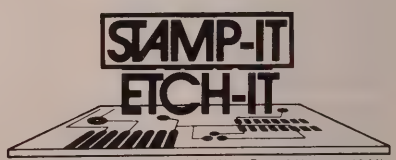
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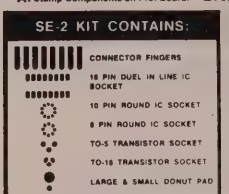


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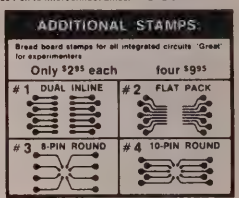


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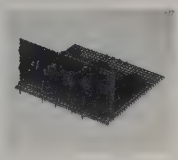
More Details? CHECK–OFF Page 110

NEW KITS ! NEW KITS !

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DIGITAL VOLTMETER



This is a 3½ digit, 0-2 volt Digital Voltmeter, with a .5% full scale accuracy. It is based around the Siliconix LD110, LD111 DVM chip set. The voltmeter uses MAN7 readouts (.3" high) to provide a highly readable display. The unit requires the following supply voltages: 12, −12, 5. The unit comes complete with all components to build the unit pictured at the left, that is a complete DVM less power supply.

$39.95 Per Kit

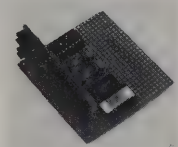
LOGIC PROBE

The Logic Probe is a unit which is for the most part indespensible in trouble shooting logic families: TTL, DTL, RTL, CMOS. It derives the power it needs to operate directly off of the circuit under test, drawing a scant 10 mA max. It uses a MAN3 readout to indicate any of the following states by these symbols: (HI)—1 (LOW)—o (PULSE)—P. The Probe can detect high frequency pulses to 45 MHz. It can't be used at MOS levels or circuit damage will result.



$9.95 Per Kit

DIGITAL COUNTER

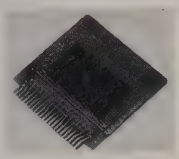


This is a 4 digit counter unit which will count up to 9999 and then provide an overflow pulse. It is based around the Mostek MK5007 digital counter chip. The unit performs the following functions: Count Input, RESET, Latch, Overflow. The counter operates up to 250 kHz. The counter is an ideal unit to be used as a frequency counter, where the only extra components needed would be a timebase, divider chain and gate. The unit requires 5V, and −12V. The unit comes complete as shown on the left less power supply.

$29.95 Per Kit

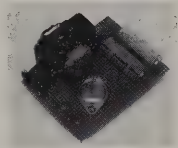
ONE KILOBYTE RANDOM ACCESS MEMORY

This memory card is for the most part a universal unit that can be used in almost any microcomputer from a HOMEBREW to an ALTAIR 8800. It uses an array of 2102 1k x 1 static random access memories to produce a 1024 x 8 memory compatible with most standard microcomputer systems. We provide everything from the super low noise vector logic card, to fine quality low profile sockets, to the eight 2102's. We even include timing diagrams and tantulumn bypass capacitor.



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5 VOLT 1 AMP T^2L SUPPLY



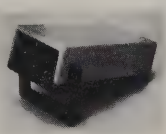
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T^2L SUPPLY

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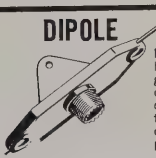
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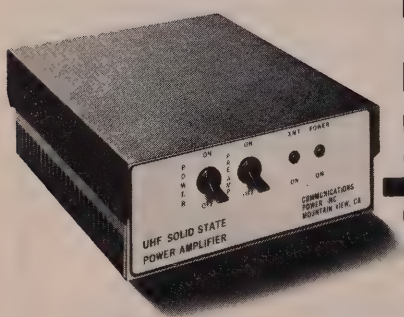
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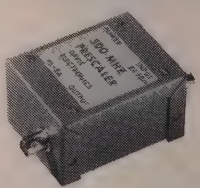
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500 MHZ PRESCALER

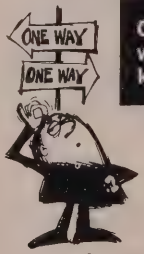
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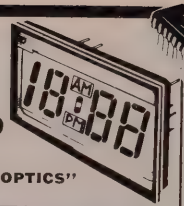
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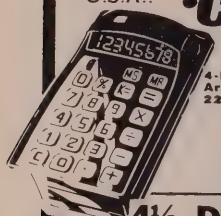
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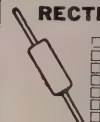
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LM-340-24T	24 v

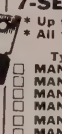
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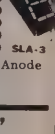




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☐	SN7425	.29	☐ SN7475	.59	☐ SN74123	.85	☐ SN74185 1.98
☐	SN7426	.25	☐ SN7476	.39	☐ SN74125	.59	☐ SN74190 1.40
☐	SN7427	.29	☐ SN7478	.79	☐ SN74126	.59	☐ SN74191 1.40
☐	SN7430	.16	☐ SN7480	.52	☐ SN74132	1.75	☐ SN74192 1.25
☐	SN7432	.25	☐ SN7481	.99	☐ SN74140	2.10	☐ SN74193 1.25
☐	SN7433	.49	☐ SN7482	.89	☐ SN74145	1.05	☐ SN74194 1.20
☐	SN7437	.34	☐ SN7483	1.25	☐ SN74148	2.25	☐ SN74195 .85
☐	SN7438	.34	☐ SN7485	1.25	☐ SN74150	.98	☐ SN74196 1.80
☐	SN7440	.16	☐ SN7486	.37	☐ SN74151	.75	☐ SN74197 .90
☐	SN7441	1.00	☐ SN7488	3.95	☐ SN74153	.90	☐ SN74199 1.75
					☐ SN74154	1.35	☐ SN74200 4.95

More Details? CHECK—OFF Page 110

flea market



RATES Non-commercial ads 10¢ per word; commercial ads 35¢ per word both payable in advance. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing). Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** can not check out each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

RTTY — Model 32 ASR - $300.00, Model 33 ASR - $600.00. Both like new and working perfectly. Can arrange shipping. K9HJV, Tony, 312-349-9002.

HOMEBREWERS: Stamp brings list of high quality components. CPO Surplus, Box 189, Braintree, Mass. 02184.

GREAT LAKES DIVISION CONVENTION — October 10 & 11, State Fairgrounds, Columbus, Ohio. Many activities planned for hams and their families: flea market forums of all types, Ohio Village tours, ceramics class, etc. $2,000 in prizes include TR4-C & two R4-C's. Advance registration $2.50. Write Mary Gibb, W8RVP, 293 Ceramic Dr., Columbus, Ohio 43214.

CANADIAN JUMBO SURPLUS and Parts Catalogs. Bargains Galore. Send $1. ETCO-HR, Box 741, Montreal "A" H3c 2V2.

ROCHESTER HAMFEST 1976 is Saturday, May 22. Your name added to mailing list or information — write: Rochester Hamfest, Box 1388, Rochester, N. Y. 14603.

SELL — Quality nylon cable ties 6 inch for $2.75 pp/hundred. New Manuals for AN/ARC-3, AN/ARC-12, AN/TRC-1 for $5.00 pp. W4VQD/Ø, 106 Sheridan Ct., Leavenworth, Ks. 66048.

FOR SALE. One of the finest amateur stations in the world. W2ONV: S-Line, linears, stacked monobanders, etc. Complete. Delivered and set up at your QTH. Inquiries - Buyers & Sellers, Box 73, Boston, Mass. 02215. 617-536-8777.

JAPANESE TRANSISTORS — All Transistors original factory made. Over 500 types available. Write for free catalog. West Pacific Electronics, P. O. Box 25837, W. Los Angeles, CA 90025.

SAROC ELEVENTH NATIONAL CONVENTION — Hotel Sahara Space Center, Las Vegas, Nevada, January 8-11, 1976. ADVANCE REGISTRATION, $12.00 per person includes: Advance and regular registration tickets; Option to purchase up to ten additional tickets for main prize drawing @$2.00 each; Admission to Friday social hour hosted by T.P.L. Communications and Tri-Ex Tower Corp., with SAROC, adults only; Admission to exhibit area and technical sessions; Admission to Saturday social hour, hosted by Ham Radio Magazine with SAROC, adults only; Hotel Sahara Safari Brunch for Sunday; Accommodations reservations card for SAROC special room rate at Hotel Sahara @$17.00 or Thunderbird Hotel @$14.00 per night, plus room tax, single or double occupancy; Tax and gratuity on all items listed except hotel room tax. ADVANCE REGISTRATION, with Midnight Show, $22.00 per person includes all items above plus Hotel Sahara's Midnight Show, with two drinks in Hotel Sahara Congo Show Room. ADVANCE REGISTRATION: with Dinner Show, $29.00 per person includes all items listed above plus Hotel Sahara's Dinner Show, Prime Rib (no drinks) in Hotel Sahara Congo Show Room. Advance registrations accepted if received in SAROC P. O. Box on or before January 1, 1976. Full refund on advance registration (after convention is over) if written request is received in SAROC P. O. Box on or before January 8. Regular registrations will be accepted at the door for an additional $2.00 each, so why not send your registration now. Special SAROC Hotel Sahara modified Safari package via most scheduled airlines serving Las Vegas from selected principal cities. Write for details. SAROC, P. O. Box 945, Boulder City, NV 89005.

PORTA-PAK the accessory that makes your mobile really portable. $59.95 and $39.95. Dealer inquiries invited. P. O. Box 67, Somers, Wisc. 53171.

VIDEO RECORDER-REPRODUCER ELECTRONIC ASSEMBLY. Contains power supply with adjustable, regulated outputs of ± 10 to 18 VDC (± 15 VDC @ 1½ amps.). Third output is 10 VDC at 3 amps. Perfect for CMOS, TTL, Op-amps. Contains over 900 useable parts with extremely long leads. Includes 182 transistors IC's, diodes, and FET's, numerous resistors, capacitors, crystals, inductors, delay lines and varicaps. Transistors will operate in HEATHKIT TV's. Schematics and semiconductor cross reference supplied upon request. Total price including shipping is $15.00. Write to Madison Electronics Company Inc., P. O. Box 369, Madison, Alabama 35758 for a free brochure.

VERY in-ter-est-ing! Next 5 big issues $1. "The Ham Trader," Sycamore IL 60178

SPRING MILL HAMFEST, Ring Mill State Park, Mitchell, IN. Sunday October 12, 1975. Advance Registrations to Hoosier Hills Ham Club, W9QYO, P. O. Box 375, Bedford, IN 47421.

QRP TRANSMATCH for HW7, Ten-Tec, and others. Send stamp for details to Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

1000V 1 AMP DIODES, 10/$1.00. Zeners, capacitors, resistors, cmos & switches for projects, free catalog. NuData Electronics, 104 N. Emerson, Mt. Prospect, IL 60056.

SOCIETY OF WIRELESS PIONEERS offers Life Membership to active and former C.W. operators on comm'l., military, gov't., etc. wireless/radio circuits. Contact: Society of Wireless Pioneers, Dept. H, P. O. Box 530, Santa Rosa, California 95402.

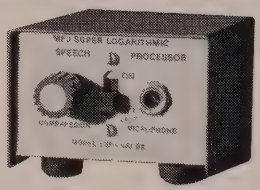
RTTY CLOSEOUT — NS-1 boards (HR 2/75) $2.75 ppd. Few units still available $29.95 ppd. Nat Stinnette Electronics, Tavares, FL 32778.

STOLEN EQUIPMENT: Standard 826M s/n 203046. Notify Jack C. Hemby, 3408 O'Hara Rd., SW., Huntsville, Ala. 35801, or the Huntsville Police Department.

BUY—SELL—TRADE. Write for monthly mailer. Give name, address, call letters. Complete stock of major brands new and reconditioned equipment. Call us for best deals. We buy Collins, Drake, Swan, etc. SSB & FM. Associated Radio, 8012 Conser, Overland Park Kansas 66204. 913-381-5901.

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NEW FROM MFJ



SUPER LOGARITHMIC SPEECH PROCESSOR

MODEL LSP-520BX

400% MORE RF POWER is yours with this plug-in unit. Simply plug **LSP-520BX** into the circuit between the microphone and transmitter and your voice suddenly is transformed from a whisper to a **DYNAMIC OUTPUT!**

Look what happens to the RF Power Output on our NCX-3. It was tuned for normal SSB operation and then left untouched for these "before" and "after" oscillograms.

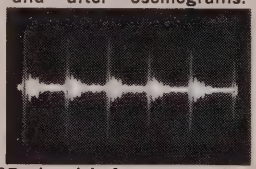


Fig. 1 SSB signal **before** processing. See the high peaks and the low valleys. Our NCX-3 is putting out only 25 watts average power.

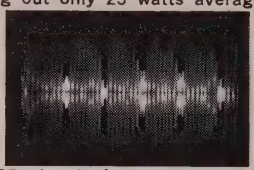


Fig. 2 SSB signal **after** processing with LSP-520BX. The once weak valleys are now strong peaks. Our NCX-3 now puts out 100 watts of average power.

Three active filters concentrate power on those frequencies that yield maximum intelligence. Adds strength in weak valleys of normal speech patterns. This is accomplished through use of an IC logarithmic amplifier with a dynamic range of 30dB for clean audio with **minimum distortion.**

This unit is practically distortion-free even at 30dB compression! The input to the LSP-520BX is completely filtered and shielded for RF protection.

Size is a mere 2 3/16H x 3½W x 4D. Money back if not delighted and **ONE YEAR UNCONDITIONAL GUARANTEE.**

Order now or write for **FREE** brochure.

LSP-520BX **$49.95**

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MFJ ENTERPRISES

P. O. BOX 494(H)

MISS. STATE, MS 39762

HAM RADIO. February 1974, wanted. Contact W1GUD, Ron Vanke, 173 Shear St. Wrentham, Mass. 02093.

NOW PAYING $2000.00 and up for ARC-94/618T ARC-102/618T. $1200.00 and up for ARC-51BX, $1500.00 and up for 490T-1 antenna couplers. We also need these control boxes — C-6287/ARC-51BX, C-6476/ARC-51BX, C-714E-2. We also need R-1051 receivers, RT-662/GRC-106 transceivers. We buy all late aircraft and ground radio equipment. Also pack radios. We are buyers not talkers. Bring your equipment in, you are paid on the spot. Ship it in, you are paid within 24 hours. We pay all shipping charges. If you want the best price for your equipment, call us. Call collect if you have, and want to sell or trade. We also sell. What do you need? D & R Electronics, R.D. #1, Box 56, Milton, Pa. 17847. Phone — 717-742-4604 - 9:00 a.m.-9:00 p.m.

CENTRAL MICHIGAN AMATEUR RADIO CLUB and the Lansing Repeater Assoc. will hold their annual Swap-Shop October 12 1975 at the Marshall Street Armory, Marshall and Saginaw St., Lansing, Mi. 9:00 a.m. to 4:00 p.m. Talk-in 34/94 and 52/52. For additional information write C.M.A.R.C., Box 73, Lansing, Mi. 48901.

RECIPROCATING DETECTOR, write Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

BOONTON 91CA RF MILLIVOLTMETER, 10 kHz to 600 MHz, 300 uv to 3 v, with hi-Z probe, $200. ME-11, A, B, C dual range wattmeter, 0-15/60w, 30-400 MHz, $65. Manuals included. James Walter, 2697 Nickel, San Pablo, Ca. 94806.

OSCAR SLIDES, set of 5, $1.25. Launch and spacecraft. Proceeds AMSAT. K6PGX, P. O. Box 463, Pasadena, CA 91102.

STILL SEARCHING for that R-390A? Let Buyers & Sellers radio brokerage do the looking for you. Buyers free; sellers 10% if you sell. All ham gear, test equipment, etc. See our display ad or write Box 73, Boston, Mass. 02215. 617-536-8777.

RECONDITIONED TEST EQUIPMENT for sale. Catalog $.50. Walter, 2697 Nickel, San Pablo, Ca. 94806.

SALE: SB401 Xmtr, $225.00; SB301 Recvr, $165.00; cords for xcvr. SB-200 linear, $215.00. SR500 Tornado xcvr SSB-CW 500 watts half gallon, $295.00. Hickok Tube checker 533A, latest chart, $75.00. Heath electrovoice mike, model 638, $29.00. Mobile mike, $7.00. Lafayette semi-automatic bug MS435, $15.00. Knight electronic VTVM 83Y125, $35.00. Leica camera, insured for $500 Leica 3F Serial 486749, 1/1000 shutter, with lens Elmar, Leitz Summaron, Summitar, valued over $100.00 each, exposure meter, flash gun, filters, many other free units. Accept best offer, or prefer swap Heath xcvrs or other. Cope, 5011 F St., Little Rock, Ark. 72205. Phone 1-501-666-7504.

CALIFORNIA QSO PARTY. October 4 & 5, sponsored by the Northern California Contest Club. The biggest party ever — all 58 of California's counties represented, 200 different W6 stations to participate. Join us.

FOR SALE: Westinghouse MW-2 R.F. assembly (transmitter), 2-30 MHz, 3 kw output. $500 or best offer. Heathkit OM-1 scope SD-3 electronic switch, $40 for both. Frank, WA9ZTO, 1721 Churchill Dr., South Bend, Ind. 46617. (219) 234-2746 after 7 p.m.

MOBILE IGNITION SHIELDING provides more range with no noise. Available most engines in assembled or kit forms, plus many other suppression accessories. Free literature. Estes Engineering, 930 Marine Dr., Port Angeles, WA. 98362.

SAVE!! Cushcraft, Hustler, Dentron, FM xtals, etc. Write for price list. Ferris Radio, 308 E. Harry, Hazel Pk., Mich. 48030. (313) 398-6645.

QSL'S — BROWNIE W3CJI — 3035B Lehigh, Allentown, Pa. 18103. Samples with cut catalog 35¢.

"HAM BUY LINES" Send name and address for Literature. Vito Iacopelli, 1720 77 St., Brooklyn, New York 11214.

WANTED: CX-7 dead or alive, write price, condition, symptoms. Sell: URM-26A, manual. Trade for 2 mtr or HF rig. WB4RSK, 2334 Regal Court, Lawrenceville, Georgia 30245.

More Details? CHECK—OFF Page 110

OHIO — The Marion Amateur Ralio Club's "Heart of Ohio HAM-FIESTA" is Sunday, October 26, at the Ohio National Guard Armory, 2 miles East of Marion on State Route 309. Advanced registration is $1.00; $1.50 at the door. For info write, Earl Adey, WB8EDO, 2697 Curren Drive, Marion, Ohio 43302.

MEMPHIS is beautiful in October! The Memphis Hamfest, bigger and better than ever, will be held at State Technical Institute, Interstate 40 at Macon Road, on Saturday and Sunday, October 4 and 5. Demonstrations, displays, MARS meetings, Flea Market, XYL entertainment, prizes. Informal dinners Saturday night. Dealers and Distributors welcome. Talk-in on 3980, .34-.94 and MARS. Contact Harry Simpson W4SCF, Box 27015, Memphis, TN 38127 or telephone 901 358-5707.

FOR SALE: Drake TR3 w/ac p.s. First certified check or money order for $260. Will ship U.P.S. where possible. Winford Barnes, Box 28, 135 W. St., Bunker Hill, IN 46914.

NEW CANADIAN MAGAZINE. "Electronics Work Shop". $5.00 yearly, sample $1.00. ETCOB, Box 741, Montreal, H3C 2V2.

FREE: 8 Extra Crystals of your choice with the purchase of a new ICOM IC-22A at $249. With the 10 crystals that come factory-installed in the IC-22A, this gives you a total of 18 crystals! For equally good deals on Kenwood, Drake, Collins, Ten-Tec, Tempo, Regency, Swan, Atlas, Midland, Alpha, ITC, Standard, Genave, Hy-Gain, Antenna Specialists, Cush-Craft, Mosley, Hustler and others, write or call Hoosier Electronics, your ham headquarters in the heart of the Midwest and become one of our many happy and satisfied customers. Hoosier Electronics, P. O. Box 2001, Terre Haute, Indiana 47802. NOTE OUR NEW PHONE NUMBER! (812)-238-1456.

TRAVEL-PAK QSL KIT — Send call and 25¢; receive your call sample kit in return. Samco, Box 203, Wynantskill, N. Y. 12198.

MANUALS for Govt. surplus gear, only $6.50 each: URM-25D, R-220/URR, SP-600JX, TT-63A/FGC, TS-497B/URR, R-274/FRR, TS-283D/U, ALR-5. Thousands more available. Send 50¢ (coin) for 22-page catalog. W3IHD, 7218 Roanne Drive, Washington, D. C. 20021.

DO-IT-URSELF DXPEDITION — Stay at ZF1SB — Cayman Is. Vertical antenna and Caribbean at your doorstep. Diving/fishing if band folds. Write Spanish Bay Reef Resort, Box 800K, Grand Cayman, B. W. I.

SALE — Mosley Classic 33 Tri-Bander, excellent $200.00; Johnson Courier 500 watt linear, $100.00; prop pitch motor, $30.00. Local - W2ETB.

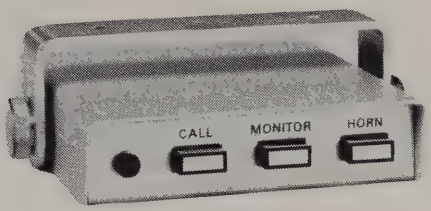
160 METER TOP LOADING SECTIONS for vertical antennas, $34.50 ppd. 80 meter sections, $31.50 ppd. Details write Bill Turney, WAØRFF, 1414 East 9th, Hutchinson, Kansas 67501.

DON' BE WITHOUT THIS new helpful 'Aid to Better DX'ing'. Taylored to fit your own Q.T.H. Over 340 countries listel. Includes amateur call prefix, country, DX zone, bearing, distance, time differential and more. Only $3.49 each. Send now to Neil Bauman, VE7ANC, 1675 Knight Avenue, Victoria, B. C., Canada V8P 1L6.

CLEVELAND'S LARGEST ELECTRONIC HAMFEST and Flea Market will be held Saturday, September 27th, at the Cuyahoga County Fairgrounds in Berea. Easy access from Hopkins Airport, Interstate I-71, I-90, and the Ohio Turnpike. Listen for the Buckeye Belles or Chix and "Talk In" from the East on 146.76, South on .82, West on .88, and locally on .52 and .94. Eastcars, Midcars on 40 and 80 M and 52.525 for farther out. Tickets $1.50 before Sept. 24th — $2.00 at 8:00 when gates open. Flea market parking earlier — bring your own tables. For early tickets and information on motels and camp grounds, send a check and S.A.S.E. to: Cleveland Hamfest Association, Box 43413, Cleveland, Ohio 44143.

VTR & CAMERA by CONCORD, EV mike, RF unit, complete like new B & W system. First money order for $150.00 takes. R. J. Morse, Box 5003, Columbia, S. C. 29250.

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MAYNARD ELECTRONICS recently purchased the entire stock of discontinued models of **PLL TOUCH-TONE DECODERS** from a major commercial manufacturer. Our agreement stipulates that the sale of these units will be limited to the amateur radio market. Because the sale price is below the manufacturer's cost, it is necessary to limit sales to five units per ham to prevent commercial speculation.

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Units are brand new. Complete satisfaction is guaranteed. If for any reason you are not completely satisfied, just return the undamaged unit within ten days for refund.

MODEL 2000B TOUCH-TONE DECODER AND MANUAL $75.00

MATING CABLE. MULTI-CONDUCTOR CABLE WITH MOLEX CONNECTOR $10.00

Please enclose check or money order for the total amount plus $5.00 for shipping and handling. A $10.00 deposit is required on C.O.D. shipments. F.O.B. Belmont. Include your name, shipping address, zip code and ham call sign. Calif. residents add sales tax.

MAYNARD ELECTRONICS COMPANY

P. O. Box 363, Dept. A
Belmont, Ca. 94002
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MM-2 MOBILE MIKE/EARPHONE	133.00
SM-3 DESK MICROPHONE	80.00
51S-1 COMMUNICATION RECEIVER	2,567.00
55G-1 PRESELECTOR	333.00
651S-1 COMMUNICATION RECEIVER	8,085.00

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P. O. BOX 100, HIGHWAY 441, DEPT. HR
OTTO, NORTH CAROLINA 28763

A Brokerage for HAM GEAR??

That's right. Now there's a place you can call to find out who's selling the equipment you're looking for—absolutely free. Sound incredible? BUYERS & SELLERS radio brokerage compiles its listings from sellers all over the country for all kinds of radio gear, test equipment, etc. If you're looking for ham gear, give us a call. We'll put you in touch with someone selling the gear you want, usually in your vicinity.

If you've got something to sell, or have some equipment that's been collecting dust, let us know about it. Sellers list their equipment at no charge and pay a 10% commission of the asking price only if we find you a buyer and a sale is made.

Do it now! Get together a list of gear you have to sell and/or would like to buy, give us the details, and we'll do the rest. When writing be sure to include your callsign, name, address, telephone, make and model of unit you're buying or selling. Sellers should also include the age, condition, price, and all serial no.s.

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BUYERS & SELLERS
Post Office Box 73
Boston, Mass. 02215

617-536-8777
Weekdays: 9am - 5pm
Wed & Sun: 7pm midnight

ANTENNA NOISE BRIDGE $19-20? Okay V.H.F., free details, Cambridge Kits, 45, Old School Lane, Milton, Cambridge, England.

FERRITE BEADS: Ferroxcube beads w/specification and application sheet — 10 @ $1.00 postpaid. Includes latest catalog. CPO Surplus, Box 189, Braintree, Mass. 02184.

FIELD STRENGTH METERS, ME-61 surplus, 1.5-24 MHz with telescoping antenna, 1½" square meter, headphone jack, range switch, all built-in. 6x6x5 aluminum box with carrying strap. Like new $12 plus $1 postage. K4CFJ, 265 Kenlock, Lexington, Kentucky 40503.

RADIO MUSEUM NOW OPEN. Free admission. 25,000 pieces of equipment from 1850 telegraph instruments to amateur and commercial transmitters of the 1920's. Amateur Station W2AN. Write for information. Antique Wireless Association, Main St., Holcomb, N. Y. 14469.

COLLINS; 30S-1 in excellent operating and physical condx for Bendix R-1051B/E receiver. Sid Sidman, 3571 Gresham Court, Pleasanton, Calif. 94566.

ENGRAVED RADIO LICENSE. Exact reproduction in solid brass. Permanent identification. Send good Xerox copy, with $5.00, to Metal Art Graphics, 1136 Potomac Ave., Hagerstown, Md. 21740.

WANTED — March 1968 thru December 1969 issues of Ham Radio. Complete sets only please. Mark Becker, 32 Oxford Rd., Newton, Mass. 02159. (617) 332-8686.

FIGHT TVI with the RSO Low Pass Filter. For brochure write: Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada. MIS 3B4.

FREE Catalog. LEDS, Microphones, Headsets, IC's, Relays, Ultrasonic Devices, Precision Trimmer Capacitors, Unique Components. Low Prices. Chaney's, Box 15431, Lakewood, Colo. 80215.

RADIO ARCHIVES, amateur ancedotes solicited for (SASE subscription) monthly PR newsletter. Electronic Avocations, 3207 4th St. N., Minneapolis, Minn. 55412.

REGENCY HR-2B in unopened carton. $190 plus $3 UPS shipping charges. Joe Trombino, 940 Alpine Road, Marion, IA 52302.

50 FREE INTERNATIONAL QSL EXCHANGE COUPONS — Send large self addressed stamped envelope, or 2 stamps. Swaney, Box 54, Goose Creek, SC. 29445.

WANTED: R-390A parts. W6ME, 4178 Chasin Street, Oceanside, Ca. 92054.

TELETYPEWRITER PARTS, gears, manuals, supplies, tape, toroids. SASE list. Typetronics, Box 8873, Ft. Lauderdale, Fl. 33310. Buy parts, late machines.

TWO PLASTIC HOLDERS frame & display 40 QSL's for $1.00, or 7 holders enhance 140 cards for $3.00. Guaranteed & prepaid, TEPABCO Box 198H, Gallatin, Tennessee 37066.

MANUALS for most ham gear made 1940/65, some earlier. Send SASE for specific quote. Hobby Industry, WØJJK, Box H-864, Council Bluffs, Iowa 51501.

5th ANNUAL SOUTHERN MINNESOTA SWAPFEST. October 11, 1975 (Saturday). Doors open 9:00 a.m. at Waseca Community High School, Waseca, Min. Southern Minnesota's largest hamfest. Contact VARS, Box 3, Waseca, Minnesota 56093. Sellers bring your own tables.

WANTED: tubes, transistors, equipment, what have you? Bernard Goldstein, W2MNP, Box 257, Canal Station, New York, N. Y. 10013.

EXCLUSIVELY HAM TELETYPE 21st year, RTTY Journal, articles, news, DX, VHF, classified ads. Sample 30¢. $3.00 per year. Box 837, Royal Oak, Michigan 48068.

PC's, Send large S.A.S.E. for list. Semtronics, Rt. #3, Box 1, Bellaire, Ohio 43906.

YOUR AD belongs here too. Commercial ads 35¢ per word. Non-commercial ads 10¢ per word. Commercial advertisers write for special discounts for standing ads not changed each month.

More Details? CHECK—OFF Page 110

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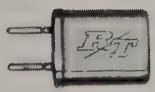


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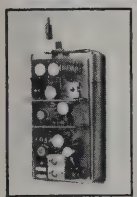
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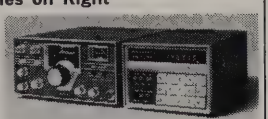
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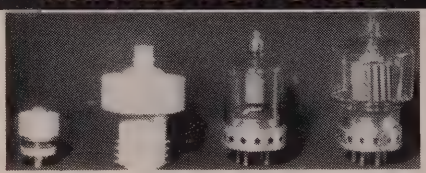
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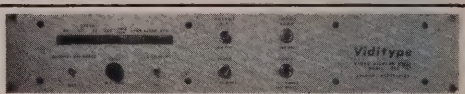
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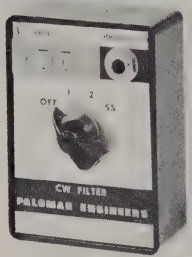
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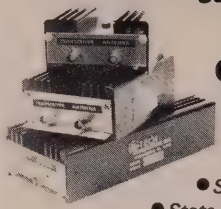


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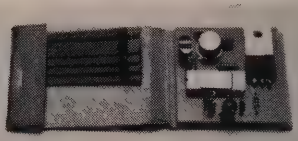
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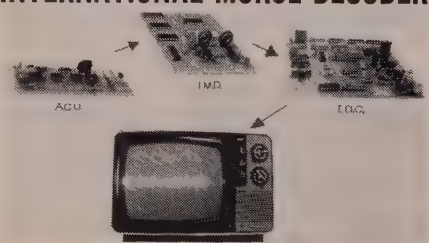
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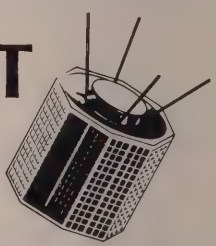
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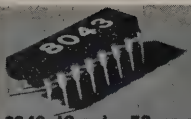
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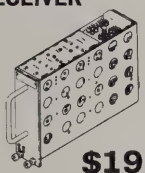
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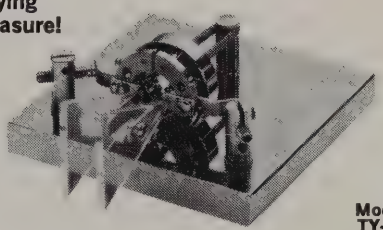
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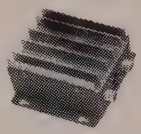
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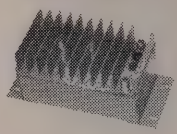
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Hy-Gain Electronics Corporation; 8601 Northeast Highway Six; Lincoln, NE 68507; 402/464-9151; Telex 48-6424. Branch Office and Warehouse; 6100 Sepulveda Blvd., #322; Van Nuys, CA 91401; 213/785-4532; Telex 65-1359. Distributed in Canada by Lectron Radio Sales, Ltd.; 211 Hunter Street West; Peterborough, Ontario.

More Details? CHECK—OFF Page 110

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. . . but you're tired of being stuck with too few channels

. . . and you'd like more versatility

. . . and you really do need tunable VFO

. . . and SSB-CW (don't forget OSCAR!)

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NEW--FROM HENRY RADIO

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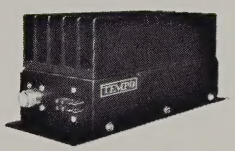
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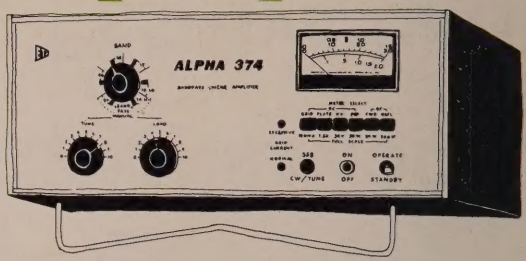
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8874-AXIAL COOLED ANODE

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